sample rate over the chip rate). For the above example in which a segment of one symbol period (e.g., N = 64 PN chips) is processed for each multipath, a buffer of two symbol periods would be able to provide a segment of one symbol period of data samples for each multipath regardless of its fractional time offset. And if the oversample rate is $N_{os} = 8$, then the minimum size of the buffer is $(2 \cdot N \cdot N_{os} = 2 \cdot 64 \cdot 8 = 1024)$ data samples.

[0092] Similarly, the capacity of the interference accumulator may be selected to be at least $(3 \cdot N \cdot N_{os})$. The extra symbol period for the interference accumulator (i.e., $3 \cdot N$ instead of $2 \cdot N$) is to account for the fact that the estimated pilot interference is derived for the next segment.

[0093] As noted above, the estimated pilot interference derived from one data sample segment may be cancelled from a later data sample segment. For a mobile terminal, the communication link and, consequently, the channel response of the various multipaths are constantly changing. Therefore, it is desirable to reduce the delay between the data samples from which the pilot interference is estimated and the data samples from which that estimated pilot interference is canceled. This delay may be as great as $2 \cdot N$ chips.

[0094] By selecting a sufficiently small value for N, the channel response of each multipath may be expected to remain relatively constant over the period of $2 \cdot N$ chips. However, the value of N should be selected to be large enough to allow for an accurate estimate of the channel response of each multipath to be processed.

[0095] FIG. 7 is a flow diagram of a process 700 to derive the total pilot interference for a number of multipaths, in accordance with an embodiment of the invention. Process 700 may be implemented by the finger processor shown in FIG. 5.

[0096] Initially, the accumulator used to accumulate the estimated pilot interferences is cleared, at step 712. An interfering multipath that has not been processed is then selected, at step 714. Typically, the pilot interference is estimated for each multipath assigned for data demodulation. However, pilot interference due to unassigned multipaths may also be estimated. In general, any number of

interfering multipaths may be processed, and these multipaths are those for which the pilot interference is to be estimated and accumulated to derive the total pilot interference.

[0097] The data samples for the received signal with the selected multipath is then processed to derive an estimate of the channel response of the selected multipath, at step 716. The channel response may be estimated based on the pilot in the selected multipath, as described above. For cdma2000, this processing entails (1) spreading the data samples with a spreading sequence for the multipath (i.e., with the proper phase corresponding to the time offset of the multipath), (2) channelizing the despread data samples to provide pilot symbols (e.g., multiplying the despread samples with the pilot channelization code and accumulating the channelized data samples over the pilot channelization code length), and (3) filtering the pilot symbols to derive pilot estimates that are indicative of the channel response of the selected multipath. Estimation of the channel response based on some other techniques may also be used, and this is within the scope of the invention.

[0098] The pilot interference due to the selected multipath is then estimated, at step 718. The pilot interference may be estimated by generating processed pilot data and multiplying this data with the estimated channel response derived in step 716. The processed pilot data is simply the spreading sequence for the selected multipath if the pilot data is a sequence of all zeros and the pilot channelization code is also all zeros. In general, the processed pilot data is the data after all signal processing at the transmitter unit but prior to the filtering and frequency upconversion (e.g., the data at the output of modulator 216a in FIG. 3 for the reverse link in cdma2000).

[0099] The estimated pilot interference for the selected multipath is then accumulated in the interference accumulator with the estimated pilot interferences for prior-processed multipaths, at step 720. As noted above, the timing phase of the multipath is observed in performing steps 716, 718, and 720.

[00100] A determination is then made whether or not all interfering multipaths have been processed, at step 722. If the answer is no, then the process returns to step 714 and another interfering multipath is selected for processing. Otherwise, the

content of the accumulator represents the total pilot interference due to all processed multipath, which may be provided in step 724. The process then terminates.

[00101] The pilot interference estimation in FIG. 7 may be performed for all multipaths in a time-division multiplexed manner using one or more finger processors. Alternatively, the pilot interference estimation for multiple multipaths may be performed in parallel using a number of finger processors. In this case, if the hardware has sufficient capabilities, then the pilot interference estimation and cancellation may be performed in real-time along with the data demodulation (e.g., as the data samples are received, with minimal or no buffering, as described above).

[00102] FIG. 8 is a flow diagram of a process 800 to data demodulate a number of multipaths with pilot interference cancellation, in accordance with an embodiment of the invention. Process 800 may also be implemented by the finger processor shown in FIG. 5.

[00103] Initially, the total pilot interference due to all multipaths of interest is derived, at step 812. Step 812 may be implemented using process 700 shown in FIG. 7. A particular multipath is then selected for data demodulation, at step 814. In an embodiment and as described above, the total pilot interference is initially canceled from the selected multipath, at step 816. This may be achieved by subtracting the interference samples for the total pilot interference (which are stored in the accumulator) from the data samples for the received signal that includes the selected multipath.

[00104] Data demodulation is then performed on the pilot-canceled signal in the normal manner. For cdma2000, this entails (1) despreading the pilot-canceled data samples, (2) channelizing the despread data to provide data symbols, and (3) demodulating the data symbols with the pilot estimates. The demodulated symbols (i.e., the demodulated data) for the selected multipath are then combined with the demodulated symbols for other multipaths for the same transmitter unit (e.g., terminal). The demodulated symbols for multipaths in multiple received signals (e.g., if receive diversity is employed) may also be combined. The symbol combining may be achieved by the symbol combiner shown in FIG. 4.

[00105] A determination is then made whether or not all assigned multipaths have been demodulated, at step 822. If the answer is no, then the process returns to step 814 and another multipath is selected for data demodulation. Otherwise, the process terminates.

[00106] As noted above, the data demodulation for all assigned multipaths of a given transmitter unit may be performed in a time-division multiplexed manner using one or more finger processors. Alternatively, the data demodulation for all assigned multipaths may be performed in parallel using a number of finger processors.

[00107] Referring back to FIGS. 4 and 5, searcher 412 may be designed and operated to search for new multipaths based on the pilot-canceled data samples (instead of the raw received data samples from buffers 408). This may provided improved search performance since the pilot interference from some or all known multipaths may have been removed as described above.

[00108] The pilot interference cancellation techniques described herein may be able to provide noticeable improvement in performance. The pilot transmitted by each terminal on the reverse link contributes to the total channel interference, I_o, in similar manner as background noise, N_o. The pilots from all terminals may represent a substantial part of the total interference level seen by all terminals. This would then result in a lower signal-to-total-noise-plus-interference ratio (SNR) for the individual terminal. In fact, it is estimated that in a cdma2000 system (which supports pilots on the reverse link) operating near capacity, approximately half of the interference seen at a base station may be due to the pilots from the transmitting terminals. Cancellation of the pilot interference may thus improve the SNR of each individual terminal, which then allows each terminal to transmit at a lower power level and increase the reverse link capacity.

[00109] The techniques described herein for estimating and canceling pilot interference may be advantageously used in various wireless communication systems that transmit a pilot along with data. For example, these techniques may be used for various CDMA systems (e.g., cdma2000, IS-95, W-CDMA, TS-CDMA, and so on), Personal Communication Services (PCS) systems (e.g., ANSI J-STD-008), and other wireless communication systems. The techniques described herein may be used to

estimate and cancel pilot interference in cases where multiple instances of each of one or more transmitted signals are received and processed (e.g., by a rake receiver or some other demodulator) and also in cases where multiple transmitted signals are received and processed.

[00110] For clarity, various aspects and embodiments of the invention have been described for the reverse link in cdma2000. The pilot interference cancellation techniques described herein may also be used for the forward link from the base station to the terminal. The processing by the demodulator is determined by the particular CDMA standard being supported and whether the inventive techniques are used for the forward or reverse link. For example, the "despreading" with a spreading sequence in IS-95 and cdma2000 is equivalent to the "descrambling" with a scrambling sequence in W-CDMA, and the channelization with a Walsh code or a quasi-orthogonal function (QOF) in IS-95 and cdma2000 is equivalent to the "despreading" with an OVSF code in W-CDMA. In general, the processing performed by the demodulator at the receiver is complementary to that performed by the modulator at the transmitter unit.

[00111] For the forward link, the techniques described herein may also be used to approximately cancel other pilots that may be transmitted in addition to, or possibly in place of, a "common" pilot transmitted to all terminals in a cell. For example, cdma2000 supports a "transmit diversity" pilot and an "auxiliary" pilot. These other pilots may utilize different Walsh codes (i.e., different channelization codes, which may be quasi-orthogonal functions). A different data pattern may also be used for the pilot. To process any of these pilots, the despread samples are decovered with the same Walsh code used to channelize the pilot at the base station, and further correlated (i.e., multiplied and accumulated) with the same pilot data pattern used at the base station for the pilot. The transmit diversity pilot and/or auxiliary pilot may be estimated and canceled in addition to the common pilot.

[00112] Similarly, W-CDMA supports a number of different pilot channels. First, a common pilot channel (CPICH) may be transmitted on a primary base station antenna. Second, a diversity CPICH may be generated based on non-zero pilot data and transmitted on a diversity antenna of the base station. Third, one or more

secondary CPICHs may be transmitted in a restricted part of the cell, and each secondary CPICH is generated using a non-zero channelization code. Fourth, the base station may further transmit a dedicated pilot to a specific user using the same channelization code as the user's data channel. In this case, the pilot symbols are time-multiplexed with the data symbols to that user. Accordingly, it will be understood by those skilled in the art that the techniques described herein are applicable for processing all of the above different types of pilot channels, and other pilot channels that may also be transmitted in a wireless communication system.

[00113] The demodulator and other processing units that may be used to implement various aspects and embodiments of the invention may be implemented in hardware, software, firmware, or a combination thereof. For a hardware design, the demodulator (including the data demodulation unit and the elements used for pilot interference estimation and cancellation such as the pilot estimator and the pilot interference estimator), and other processing units may be implemented within one or more application specific integrated circuits (ASIC), digital signal processors (DSP), digital signal processing devices (DSPDs), field programmable gate arrays (FPGA), processors, microprocessors, controllers, microcontrollers, programmable logic devices (PLD), other electronic units, or any combination thereof.

[00114] For a software implementation, the elements used for pilot interference estimation and cancellation and data demodulation may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory 262 in FIG. 2) and executed by a processor (e.g., controller 260). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as it known in the art.

[00115] The elements used to implement the pilot interference estimation and cancellation described herein may be incorporated in a receiver unit or a demodulator that may further be incorporated in a terminal (e.g., a handset, a handheld unit, a stand-alone unit, and so on), a base station, or some other communication devices or units. The receiver unit or demodulator may be implemented with one or more integrated circuits.

[00116] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WHAT IS CLAIMED IS:

1. A method for canceling pilot interference at a receiver unit in a wireless communication system, comprising:

receiving a signal comprised of a plurality of signal instances, wherein each signal instance includes a pilot;

deriving total pilot interference due to one or more signal instances;

subtracting the total pilot interference from the received signal to derive a pilot-canceled signal; and

processing the pilot-canceled signal to derive demodulated data for each of at least one signal instance in the received signal.

2. The method of claim 1, wherein the total pilot interference is derived by

estimating pilot interference due to each of the one or more signal instances, and

accumulating the estimated pilot interference for the one or more signal instances.

3. The method of claim 2, wherein the pilot interference due to each of the one or more signal instances is estimated by

processing the signal instance to derive an estimate of a channel response of the signal instance, and

multiplying processed pilot data for the signal instance with the estimated channel response to provide the estimated pilot interference.

4. The method of claim 3, wherein the processed pilot data for each of the one or more signal instances is a spreading sequence for the signal instance.

5. The method of claim 4, wherein the spreading sequence for the signal instance has a phase corresponding to an arrival time of the signal instance.

6. The method of claim 3, wherein the estimated channel response for each of the one or more signal instances is derived by

despreading data samples for the received signal with a spreading sequence for the signal instance,

channelizing the despread samples with a pilot channelization code to provide pilot symbols, and

filtering the pilot symbols to provide the estimated channel response.

7. The method of claim 3, wherein the estimated channel response of the signal instance is derived based on a current segment of data samples for the received signal and the estimated pilot interference is for a subsequent segment of data samples.

8. The method of claim 3, wherein the estimated channel response of the signal instance is derived based on a current segment of data samples for the received signal and the estimated pilot interference is for the same segment of data samples.

9. The method of claim 3, wherein the estimated channel response for each of the one or more signal instances is derived based on data samples for the received signal.

10. The method of claim 3, wherein the estimated channel response for each of the one or more signal instances is derived based on data samples having pilot from the signal instance unremoved but pilots from other interfering signal instances removed.

11. The method of claim 1, wherein the processing of the pilot-canceled signal for each of the at least one signal instance includes

despreading samples for the pilot-canceled signal with a spreading sequence for the signal instance,

channelizing the despread samples with a data channelization code to provide data symbols, and

demodulating the data symbols with pilot estimates to provide the demodulated data for the signal instance.

12. The method of claim 11, wherein the pilot estimates for each of the at least one signal instance are derived based on data samples for the received signal.

13. The method of claim 11, wherein the pilot estimates for each of the at least one signal instance are derived based on data samples having pilot from the signal instance unremoved but pilots from other interfering signal instances removed.

14. The method of claim 2, wherein the pilot interference due to the one or more signal instances is estimated in a time-division multiplexed manner.

15. The method of claim 1, wherein the subtracting includes

subtracting interference samples for the total pilot interference from data samples for the received signal, wherein the interference samples and data samples are both provided at a particular sample rate.

16. The method of claim 1, wherein the pilot interference due to a signal instance being processed to derive the demodulated data is excluded from the total pilot interference.

17. The method of claim 1, further comprising:

processing the pilot-canceled signal to search for new signal instances in the received signal.

18. The method of claim 15, wherein the sample rate is multiple times a chip rate.

19. The method of claim 1, wherein the deriving the total pilot interference is performed based on segments of data samples for the received signal.

20. The method of claim 19, wherein the each segment includes data samples for one symbol period.

21. The method of claim 1, wherein the processing to derive demodulated data is performed based on segments of pilot-canceled data samples for the pilot-canceled signal.

22. The method of claim 1, wherein the deriving the total pilot interference and the processing of the pilot-canceled signal are performed in parallel.

23. The method of claim 1, wherein the deriving the total pilot interference and the processing of the pilot-canceled signal are performed in a pipelined manner.

24. The method of claim 1, wherein the wireless communication system is a CDMA system.

25. The method of claim 24, wherein the CDMA system supports cdma2000 standard.

26. The method of claim 24, wherein the CDMA system supports W-CDMA standard.

27. The method of claim 24, wherein the CDMA system supports IS-95 standard.

28. The method of claim 24, wherein the received signal comprises one or more reverse link modulated signals in the CDMA system.

29. The method of claim 24, wherein the received signal comprises one or more forward link modulated signals in the CDMA system.

30. A method for canceling pilot interference at a receiver unit in a wireless communication system, comprising:

processing a received signal comprised of a plurality of signal instances to provide data samples, wherein each signal instance includes a pilot;

processing the data samples to derive an estimate of pilot interference due to each of one or more signal instances;

deriving total pilot interference due to the one or more signal instances based on the estimated pilot interference;

subtracting the total pilot interference from the data samples to derive pilot-canceled data samples; and

processing the pilot-canceled data samples to derive demodulated data for each of at least one signal instance in the received signal.

31. The method of claim 30, wherein the processing the data samples to derive the estimated pilot interference due to each of the one or more signal instances includes

despreading the data samples with a spreading sequence for the signal instance,

channelizing the despread samples with a pilot channelization code to provide pilot symbols,

filtering the pilot symbols to provide an estimate or a channel response of the signal instance, and

multiplying the spreading sequence for the signal instance with the estimated channel response to provide the estimated pilot interference due to the signal instance.

32. The method of claim 30, wherein the processing the pilot-canceled data samples to derive the demodulated data for each of the at least one signal instance includes

despreading the pilot-canceled data samples with a spreading sequence for the signal instance,

channelizing the despread samples with a data channelization code to provide data symbols, and

demodulating the data symbols to provide the demodulated data for the signal instance.

33. The method of claim 30, wherein the subtracting includes

subtracting interference samples for the total pilot interference from the data samples for the received signal, wherein the interference samples

and data samples are both provided at a particular sample rate that is multiple times a chip rate.

34. A receiver unit in a wireless communication system, comprising:

a receiver configured to process a received signal comprised of a plurality of signal instances to provide data samples, wherein each signal instance includes a pilot; and

a demodulator including

a pilot interference estimator configured to process the data samples to derive an estimate of pilot interference due to each of one or more signal instances and to derive total pilot interference due to the one or more signal instances based on the estimated pilot interference,

a summer configured to subtract the total pilot interference from the data samples to derive pilot-canceled data samples, and

a data demodulation unit configured to process the pilot-canceled data samples to derive demodulated data for each of at least one signal instance in the received signal.

35. The receiver unit of claim 34, wherein the demodulator further includes

a channel estimator configured to provide an estimated channel response for each of the one or more signal instances.

36. The receiver unit of claim 35, wherein the pilot interference estimator is further configured to multiply processed pilot data for each of the one or more signal instances with the estimated channel response for the signal instance to provide the estimated pilot interference due to the signal instance.

37. The receiver unit of claim 34, wherein for each of the at least one signal instance the data demodulation unit is configured to despread the pilot-canceled data samples with a spreading sequence for the signal instance, channelize the despread samples with a data channelization code to provide data symbols, and demodulate the data symbols with pilot estimates for the signal instance to provide the demodulated data for the signal instance.

38. A terminal in a CDMA system comprising:

a receiver configured to process a received signal comprised of a plurality of signal instances to provide data samples, wherein each signal instance includes a pilot; and

a demodulator including

a pilot interference estimator configured to process the data samples to derive an estimate of pilot interference due to each of one or more signal instances and to derive total pilot interference due to the one or more signal instances based on the estimated pilot interference,

a summer configured to subtract the total pilot interference from the data samples to derive pilot-canceled data samples, and

a data demodulation unit configured to process the pilot-canceled data samples to derive demodulated data for each of at least one signal instance in the received signal.

39. The terminal of claim 38, wherein the demodulator further includes

a channel estimator configured to provide an estimated channel response for each of the one or more signal instances.

40. The terminal of claim 39, wherein the pilot interference estimator is further configured to multiply processed pilot data for each of the one or

more signal instances with the estimated channel response for the signal instance to provide the estimated pilot interference due to the signal instance.

41. The terminal of claim 38, wherein for each of the at least one signal instance the data demodulation unit is configured to despread the pilot-canceled data samples with a spreading sequence for the signal instance, channelize the despread samples with a data channelization code to provide data symbols, and demodulate the data symbols with pilot estimates for the signal instance to provide the demodulated data for the signal instance.

42. A base station in a CDMA system comprising:

a receiver configured to process a received signal comprised of a plurality of signal instances to provide data samples, wherein each signal instance includes a pilot; and

a demodulator including

a pilot interference estimator configured to process the data samples to derive an estimate of pilot interference due to each of one or more signal instances and to derive total pilot interference due to the one or more signal instances based on the estimated pilot interference,

a summer configured to subtract the total pilot interference from the data samples to derive pilot-canceled data samples, and

a data demodulation unit configured to process the pilot-canceled data samples to derive demodulated data for each of at least one signal instance in the received signal.

43. The base station of claim 42, wherein the demodulator further includes

a channel estimator configured to provide an estimated channel response for each of the one or more signal instances.

44. The base station of claim 43, wherein the pilot interference estimator is further configured to multiply processed pilot data for each of the one or more signal instances with the estimated channel response for the signal instance to provide the estimated pilot interference due to the signal instance.

45. The base station of claim 42, wherein for each of the at least one signal instance the data demodulation unit is configured to despread the pilot-canceled data samples with a spreading sequence for the signal instance, channelize the despread samples with a data channelization code to provide data symbols, and demodulate the data symbols with pilot estimates for the signal instance to provide the demodulated data for the signal instance.

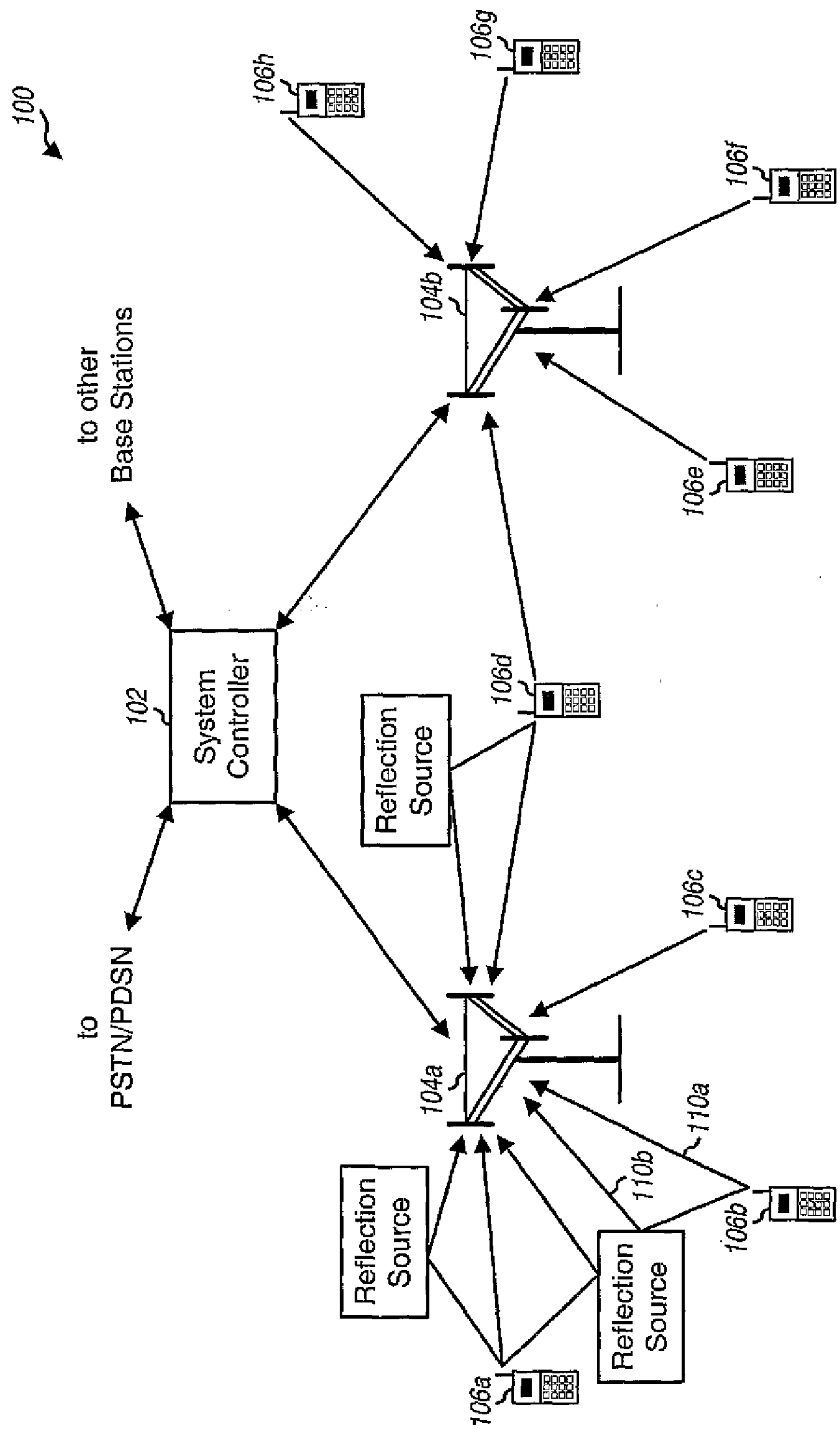

100
to other
Base Stations
to
PSTN/PDSN
102
System
Controller
104a
104b
106a
106b
106c
106d
106e
106f
106g
106h
110a
110b
Reflection
Source
Reflection
Source
Reflection
Source

FIG. 1

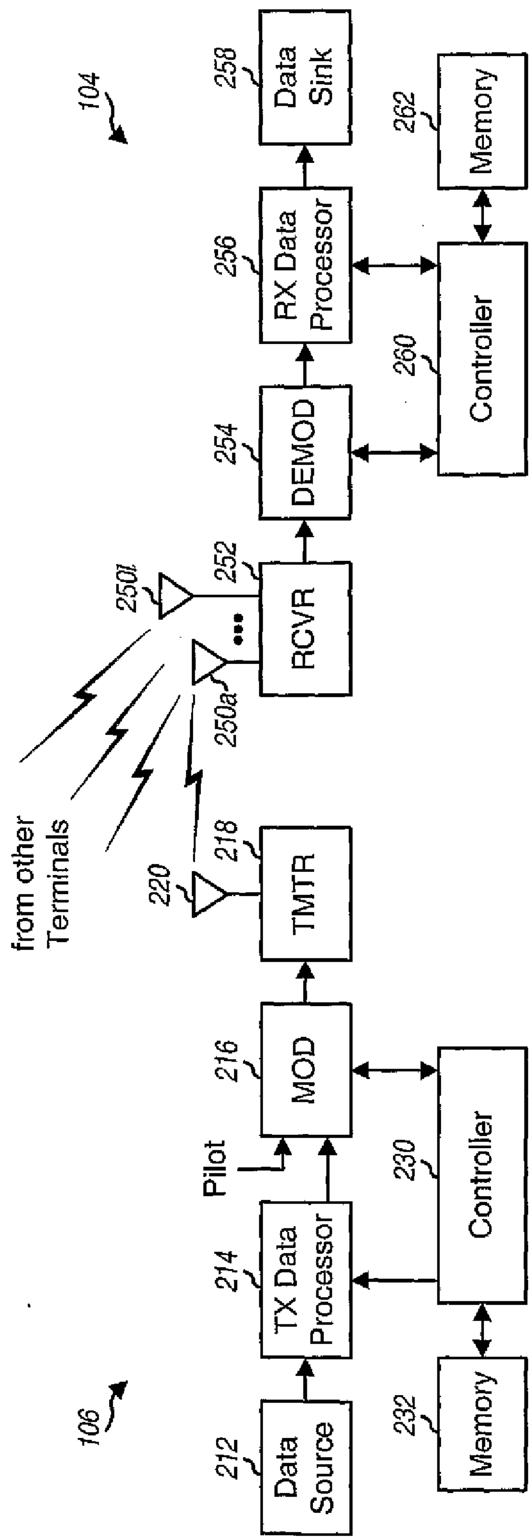

104
258
Data Sink
256
RX Data Processor
254
DEMOD
252
RCVR
250l
250a
from other Terminals
262
Memory
260
Controller
218
TMTR
220
216
MOD
Pilot
214
TX Data Processor
230
Controller
106
212
Data Source
232
Memory

FIG. 2

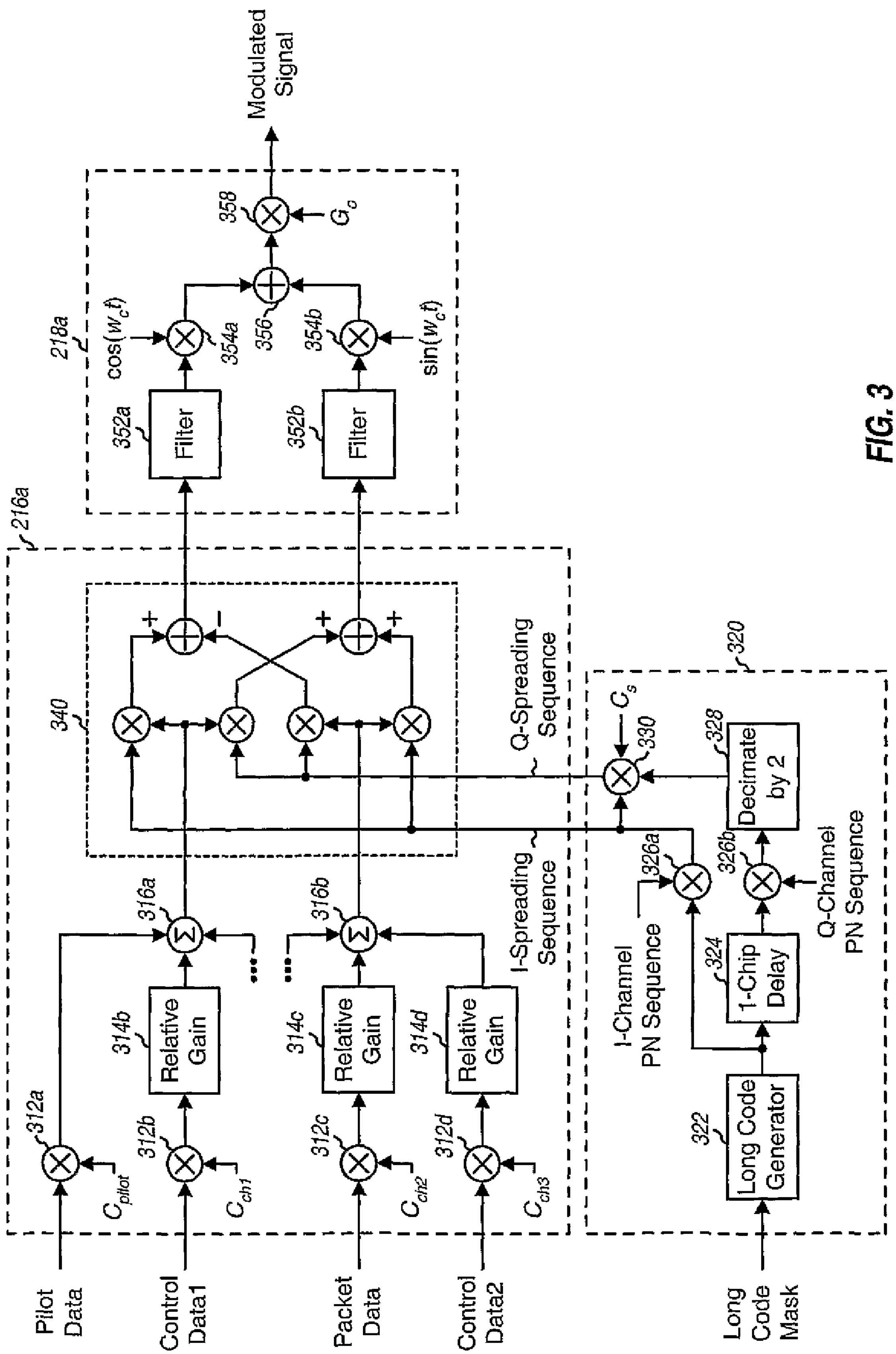

Pilot Data
Control Data1
Packet Data
Control Data2
Long Code Mask
312a
312b
312c
312d
C pilot
C ch1
C ch2
C ch3
314b
314c
314d
Relative Gain
Relative Gain
Relative Gain
316a
316b
Σ
Σ
216a
340
I-Spreading Sequence
Q-Spreading Sequence
320
322
Long Code Generator
324
1-Chip Delay
326a
326b
328
Decimate by 2
330
C s
I-Channel PN Sequence
Q-Channel PN Sequence
218a
352a
352b
Filter
Filter
354a
354b
cos(w c t)
sin(w c t)
356
358
G o
Modulated Signal

FIG. 3

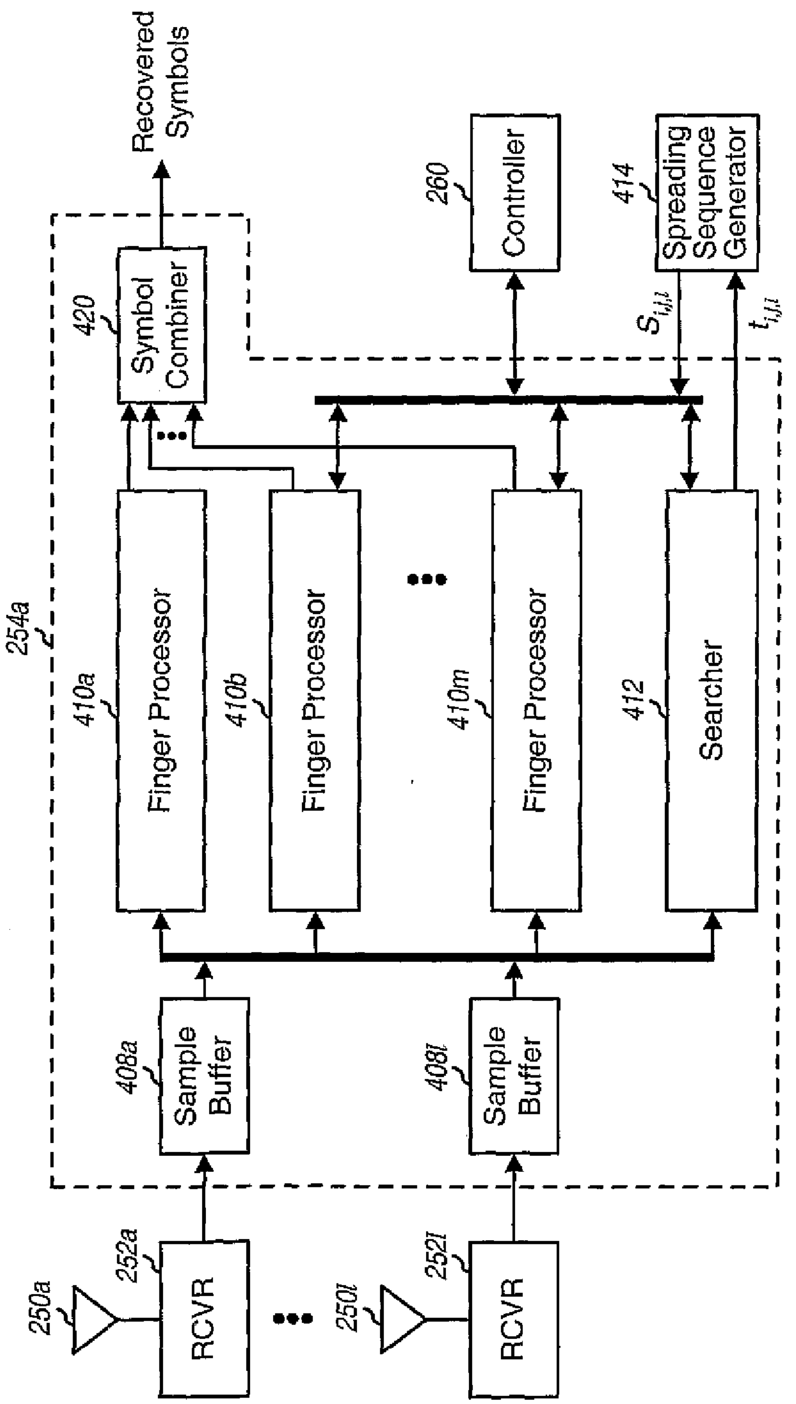
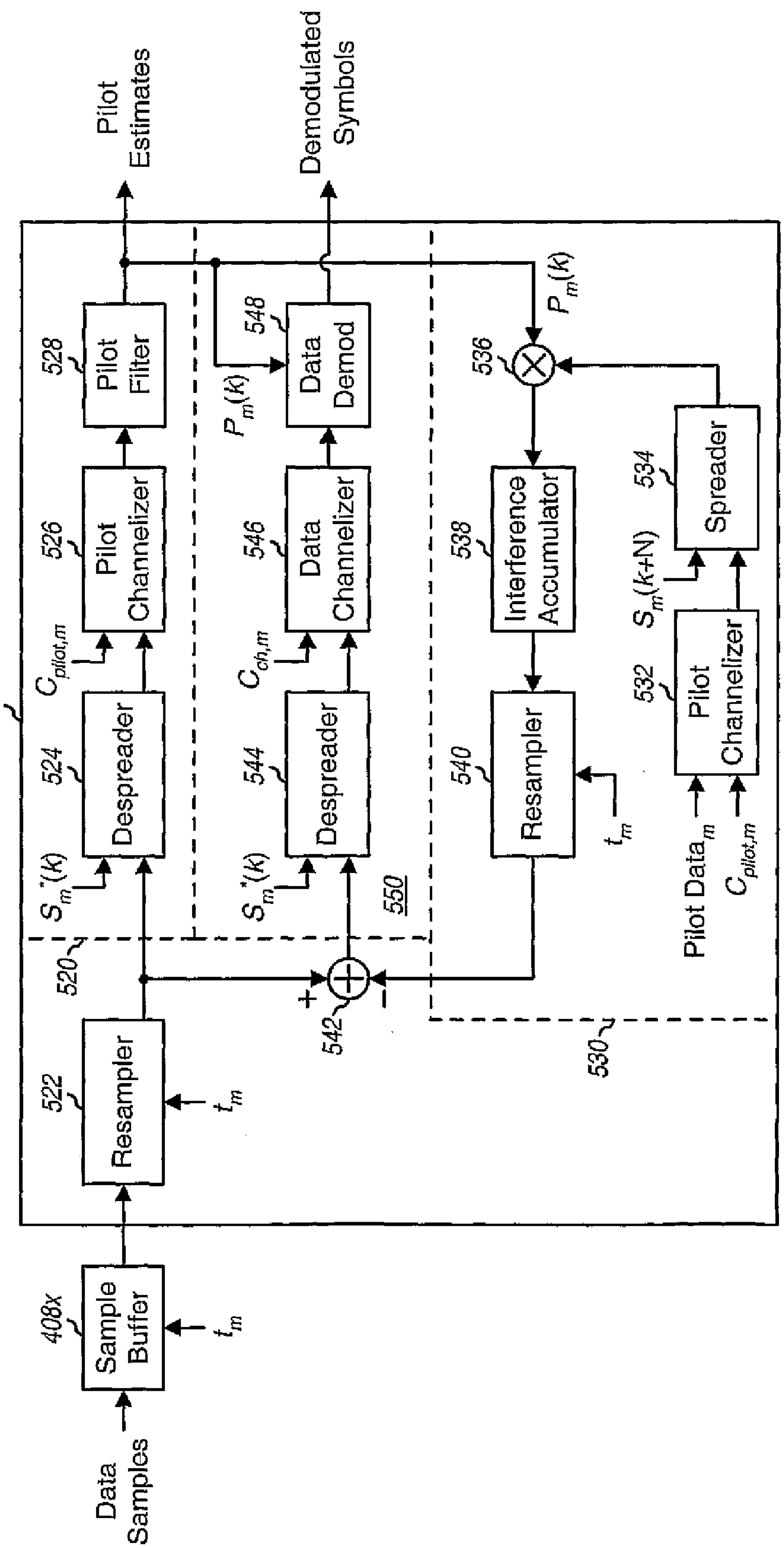

Recovered Symbols
420
Symbol Combiner
260
Controller
414
Spreading Sequence Generator
$S_{i,j,l}$
$t_{i,j,l}$
254a
410a
Finger Processor
410b
Finger Processor
410m
Finger Processor
412
Searcher
408a
Sample Buffer
408l
Sample Buffer
250a
252a
RCVR
250l
252l
RCVR

FIG. 4



Data Samples
408x
Sample Buffer
t_m
410x
522
Resampler
t_m
520
542
+
−
530
524
Despreader
$S_m^*(k)$
$C_{pilot,m}$
526
Pilot Channelizer
528
Pilot Filter
Pilot Estimates
544
Despreader
$S_m^*(k)$
550
$C_{ch,m}$
546
Data Channelizer
548
Data Demod
$P_m(k)$
Demodulated Symbols
536
$P_m(k)$
538
Interference Accumulator
540
Resampler
t_m
532
Pilot Channelizer
Pilot $Data_m$
$C_{pilot,m}$
$S_m(k+N)$
534
Spreader

FIG. 5

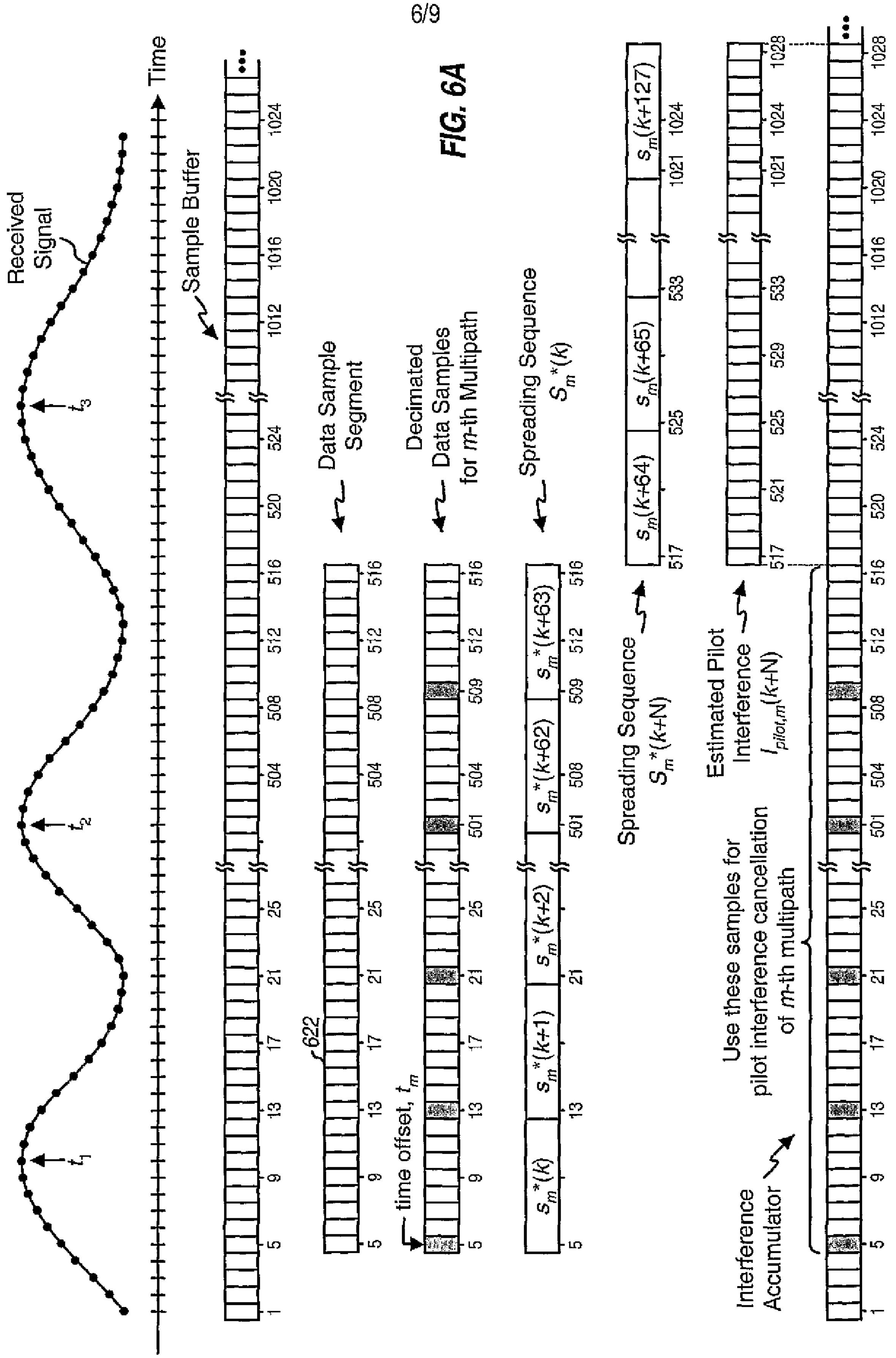

Received Signal
Time
t_1
t_2
t_3
Sample Buffer
Data Sample Segment
622
Decimated Data Samples for m-th Multipath
time offset, t_m
Spreading Sequence $S_m{}^*(k)$
$s_m{}^*(k)$
$s_m{}^*(k+1)$
$s_m{}^*(k+2)$
$s_m{}^*(k+62)$
$s_m{}^*(k+63)$
Spreading Sequence $S_m{}^*(k+N)$
$s_m(k+64)$
$s_m(k+65)$
$s_m(k+127)$
Estimated Pilot Interference $I_{pilot,m}(k+N)$
Interference Accumulator
Use these samples for pilot interference cancellation of m-th multipath

FIG. 6A

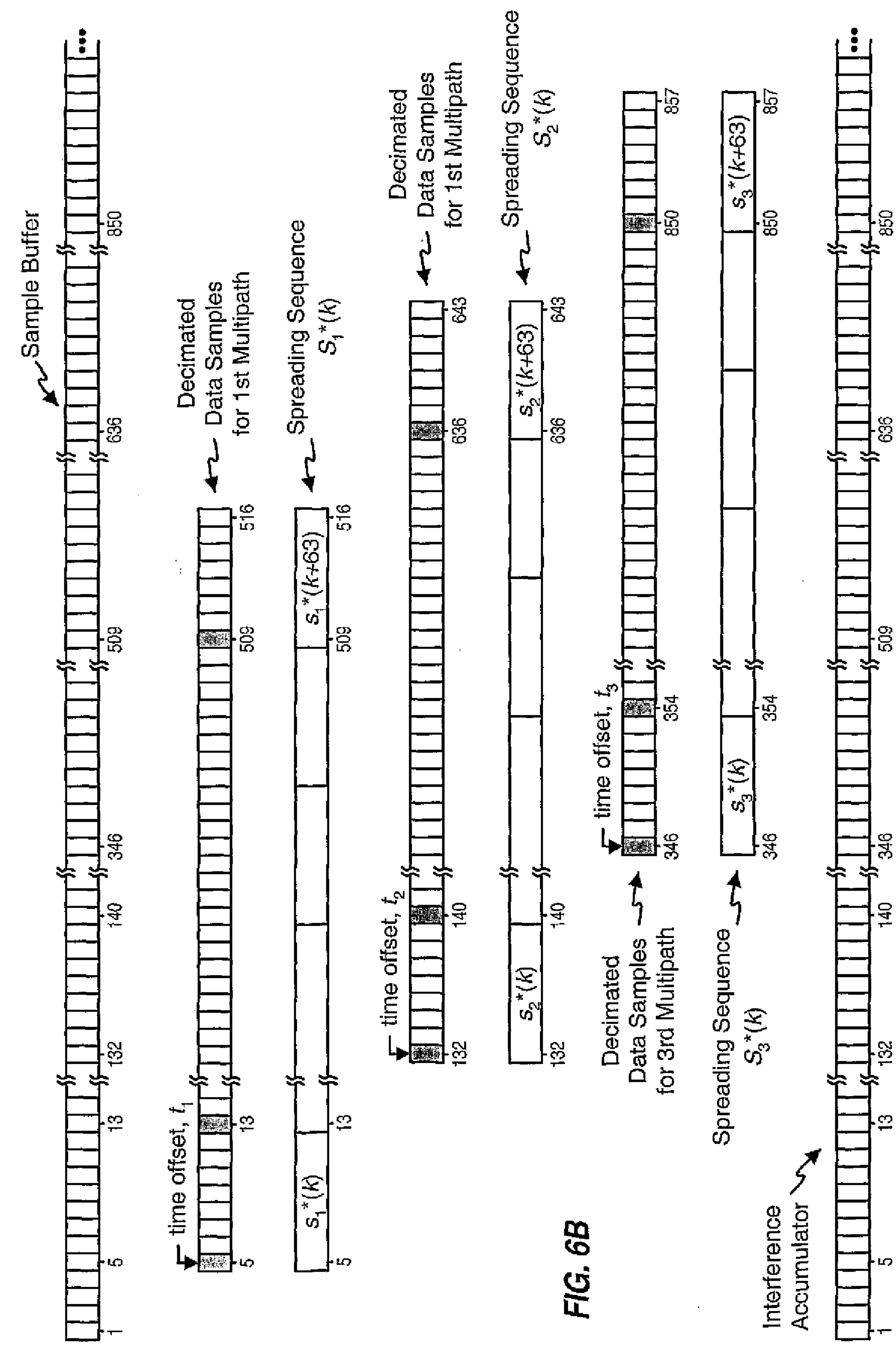

Sample Buffer
1
5
13
132
140
346
509
636
850
time offset, t_1
Decimated
Data Samples
for 1st Multipath
516
Spreading Sequence
$S_1^*(k)$
$s_1^*(k)$
$s_1^*(k+63)$
time offset, t_2
Decimated
Data Samples
for 1st Multipath
643
Spreading Sequence
$S_2^*(k)$
$s_2^*(k)$
$s_2^*(k+63)$
time offset, t_3
Decimated
Data Samples
for 3rd Multipath
354
857
Spreading Sequence
$S_3^*(k)$
$s_3^*(k)$
$s_3^*(k+63)$
Interference
Accumulator

FIG. 6B

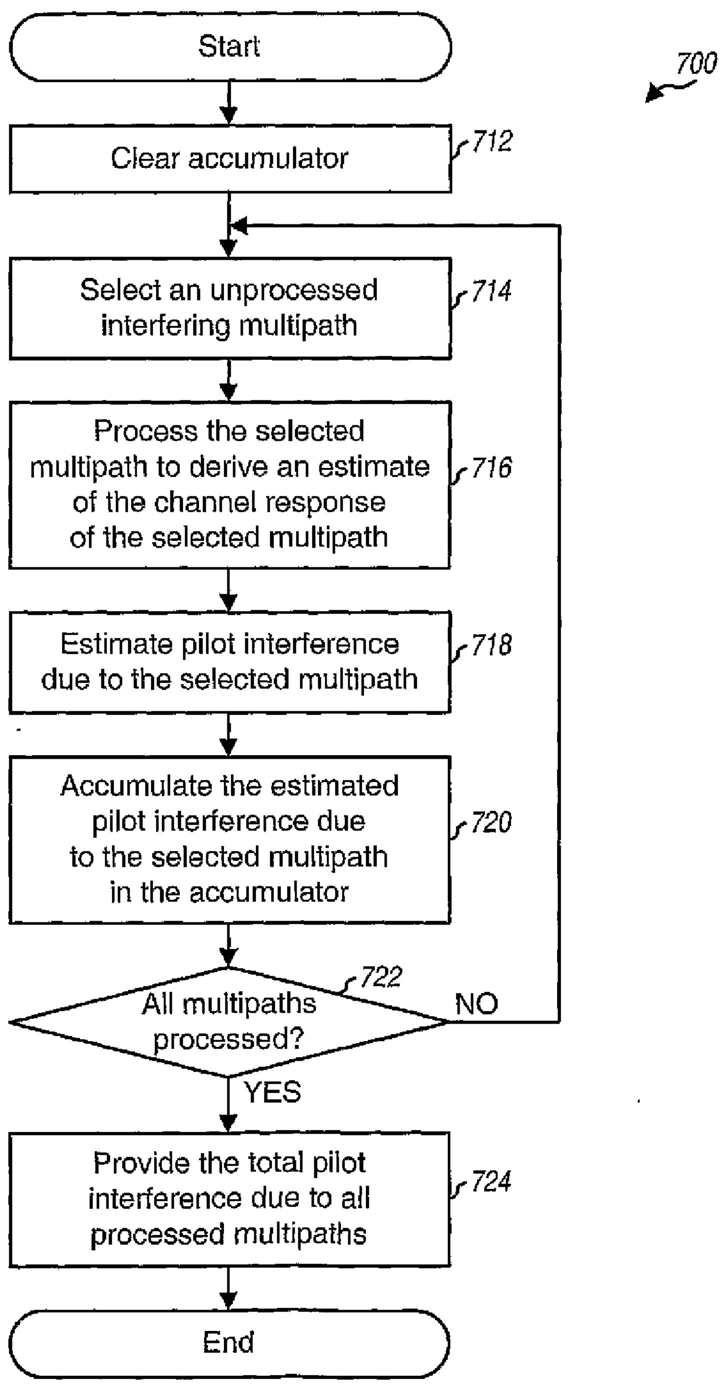

Start
700
Clear accumulator
712
Select an unprocessed interfering multipath
714
Process the selected multipath to derive an estimate of the channel response of the selected multipath
716
Estimate pilot interference due to the selected multipath
718
Accumulate the estimated pilot interference due to the selected multipath in the accumulator
720
All multipaths processed?
722
NO
YES
Provide the total pilot interference due to all processed multipaths
724
End

FIG. 7

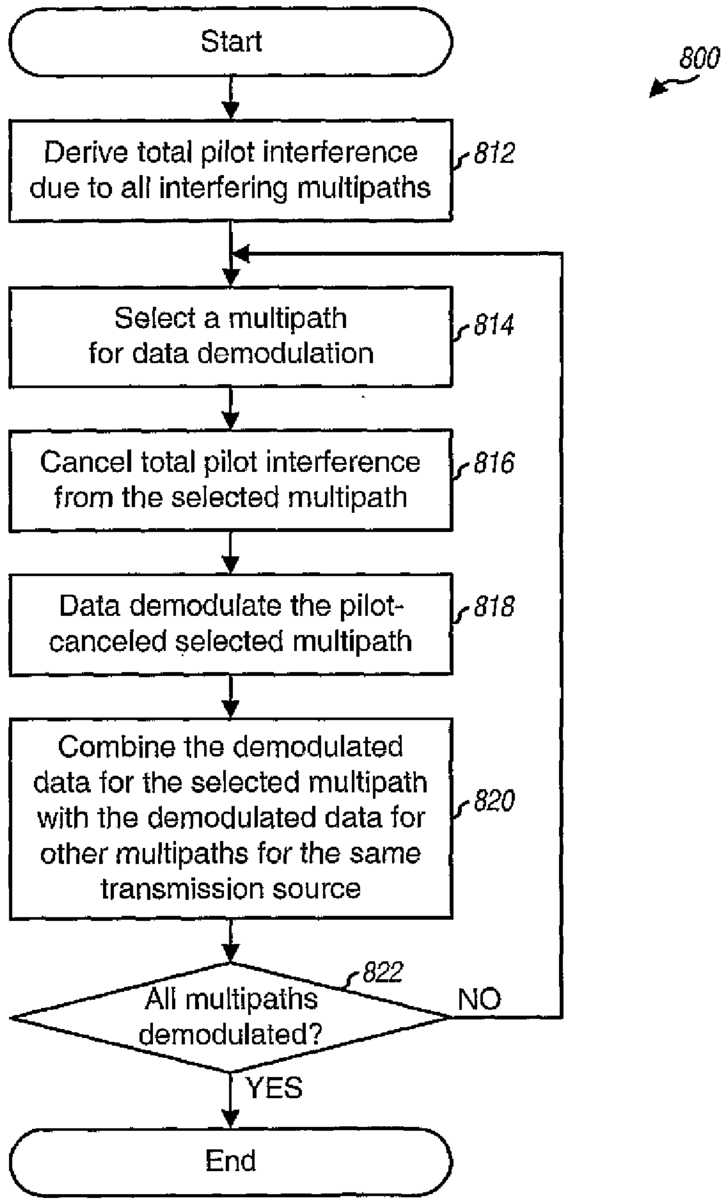

Start
800
Derive total pilot interference due to all interfering multipaths
812
Select a multipath for data demodulation
814
Cancel total pilot interference from the selected multipath
816
Data demodulate the pilot-canceled selected multipath
818
Combine the demodulated data for the selected multipath with the demodulated data for other multipaths for the same transmission source
820
All multipaths demodulated?
822
NO
YES
End

FIG. 8

INTERNATIONAL SEARCH REPORT

Inte   il Application No
PCI/US 02/18133

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04B1/707

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

INSPEC, EPO-Internal

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 6 067 292 A (BRINK STEPHAN TEN ET AL) 23 May 2000 (2000-05-23) abstract; figures 6,7,10,12,14,16,16S,17,22,23 column 2, line 1 - line 42 column 7, line 4 -column 12, line 28 column 15, line 10 - line 22 ---	1-5

-/--

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
19 September 2002	30/09/2002

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Bauer, F

Int al Application No	PCT/US 02/18133

INTERNATIONAL SEARCH REPORT

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	IWAKIRI N: "INTERFERENCE REDUCTION EFFICIENCY OF A TURBO CODED CDMA MULTILAYER SYSTEM EQUIPPED WITH A PILOT CANCELER" VTC 1999-FALL. IEEE VTS 50TH. VEHICULAR TECHNOLOGY CONFERENCE. GATEWAY TO THE 21ST. CENTURY COMMUNICATIONS VILLAGE. AMSTERDAM, SEPT. 19 - 22, 1999, IEEE VEHICULAR TECHNOLGY CONFERENCE, NEW YORK, NY: IEEE, US, vol. 1 CONF. 50, September 1999 (1999-09), pages 391-395, XP000929078 ISBN: 0-7803-5436-2 paragraph '000C!; figure 3 ---	1
X	EP 0 980 149 A (IND TECH RES INST) 16 February 2000 (2000-02-16) paragraphs '0011!,'0017!,'0026!-'0032!; figure 2A -----	1-5

INTERNATIONAL SEARCH REPORT

tional application No.
PCT/US 02/18133

Box I Observations where certain claims were found unsearchable (Continuation of item 1 of first sheet)

This International Search Report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. [ ] Claims Nos.:
because they relate to subject matter not required to be searched by this Authority, namely:

2. [X] Claims Nos.: 6-45
because they relate to parts of the International Application that do not comply with the prescribed requirements to such an extent that no meaningful International Search can be carried out, specifically:
see FURTHER INFORMATION sheet PCT/ISA/210

3. [ ] Claims Nos.:
because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a).

Box II Observations where unity of invention is lacking (Continuation of item 2 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

1. [ ] As all required additional search fees were timely paid by the applicant, this International Search Report covers all searchable claims.

2. [ ] As all searchable claims could be searched without effort justifying an additional fee, this Authority did not invite payment of any additional fee.

3. [ ] As only some of the required additional search fees were timely paid by the applicant, this International Search Report covers only those claims for which fees were paid, specifically claims Nos.:

4. [ ] No required additional search fees were timely paid by the applicant. Consequently, this International Search Report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

Remark on Protest

[ ] The additional search fees were accompanied by the applicant's protest.

[ ] No protest accompanied the payment of additional search fees.

FURTHER INFORMATION CONTINUED FROM PCT/ISA/ 210

Continuation of Box I.2

Claims Nos.: 6-45

In view of the large number of independent claims (5), and on the even larger number of claims dependent on the not novel calims 1 and 3 (16) , which render it difficult, if not impossible, to determine the matter for which protection is sought, the present application fails to comply with the clarity and conciseness requirements of Article 6 PCT (see also Rule 6.1(a) PCT) to such an extent that a meaningful search is impossible. Consequently, the search has been carried out for those parts of the application which do appear to be clear (and concise), namely claims 1-5.

The applicant's attention is drawn to the fact that claims, or parts of claims, relating to inventions in respect of which no international search report has been established need not be the subject of an international preliminary examination (Rule 66.1(e) PCT). The applicant is advised that the EPO policy when acting as an International Preliminary Examining Authority is normally not to carry out a preliminary examination on matter which has not been searched. This is the case irrespective of whether or not the claims are amended following receipt of the search report or during any Chapter II procedure.

INTERNATIONAL SEARCH REPORT

Information on patent family members

Int l Application No
PCT/US 02/18133

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
US 6067292	A	23-05-2000	US	6009089 A	28-12-1999
			EP	0876002 A2	04-11-1998
			JP	10327126 A	08-12-1998
			KR	263801 B1	16-08-2000
EP 0980149	A	16-02-2000	US	6154443 A	28-11-2000
			EP	0980149 A2	16-02-2000
			TW	419912 B	21-01-2001

ORTHOGONAL FREQUENCY-DIVISION MULTIPLEX TRANSMISSION METHOD

Publication number: WO03047140 (A1)
Publication date: 2003-06-05
Inventor(s): NAKAMURA TAKAHARU [JP] +
Applicant(s): FUJITSU LTD [JP]; NAKAMURA TAKAHARU [JP] +
Classification:
- international: ***H04B7/005; H04J11/00; H04L27/26; H04L5/02;*** **H04B7/005; H04J11/00; H04L27/26; H04L5/02;** (IPC1-7): H04J11/00
- European: H04J11/00; H04L27/26M1G; H04L5/00A3A; H04W52/26R; H04W52/28
Application number: WO2001JP10357 20011128
Priority number(s): WO2001JP10357 20011128

Also published as:

EP1450505 (A1)
EP1450505 (A4)
EP1450505 (B1)
US2004213145 (A1)
US2007183310 (A1)

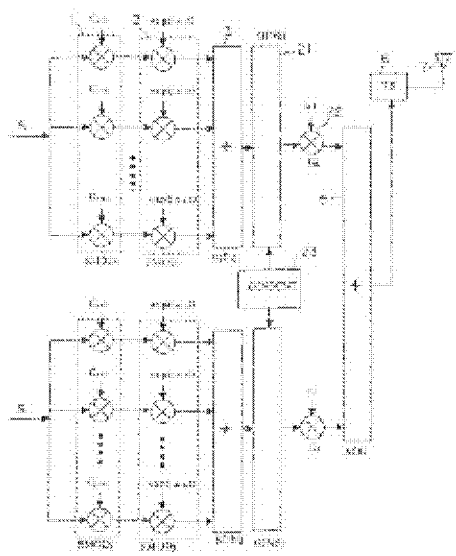
more >>

Cited documents:

EP1035693 (A2)
JP2000165342 (A)
EP1014639 (A2)
JP2001111519 (A)
JP11196062 (A)

Abstract of **WO 03047140 (A1)**

A spreading modulator 1 spreads the spectrum of a signal series. A subcarrier modulator 2 modulates the frequencies of a plurality of subcarriers having frequencies different from one another by using the output of the spreading modulator 1. An adder 3 combines the subcarriers modulated. A guard section control unit 23 determines the length of a guard section in accordance with the maximum transmission delay difference of a line between a transmission device and a reception device. A guard section inserter 21 is controlled by the guard section control unit 23 to insert the guard section into the signal series for every symbol period. A gain adjustor 22 multiplies the transmission signal by a gain coefficient corresponding to the guard section inserted.A spreading modulator (1) spreads the spectrum of a signal series. A subcarrier modulator (2) modulates the frequencies of a plurality of subcarriers having frequencies different from one another by using the output of the spreading modulator (1). An adder (3) combines the subcarriers modulated. A guard section control unit (23) determines the length of a guard section in accordance with the maximum transmission delay difference of a line between a transmission device and a reception device. A guard section inserter (21) is controlled by the guard section control unit (23) to insert the guard section into the signal series for every symbol period. A gain adjustor (22) multiplies the transmission signal by a gain coefficient corresponding to the guard section inserted.



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(19) 世界知的所有権機関
国際事務局

(43) 国際公開日
2003 年6 月5 日 (05.06.2003)

PCT

(10) 国際公開番号
WO 03/047140 A1

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(21) 国際出願番号: PCT/JP01/10357

(22) 国際出願日: 2001 年11 月28 日 (28.11.2001)

(25) 国際出願の言語: 日本語

(26) 国際公開の言語: 日本語

(71) 出願人 (米国を除く全ての指定国について): 富士通株式会社 (FUJITSU LIMITED) [JP/JP]; 〒211-8588 神奈川県川崎市中原区上小田中4丁目1番1号 Kanagawa (JP).

(72) 発明者; および

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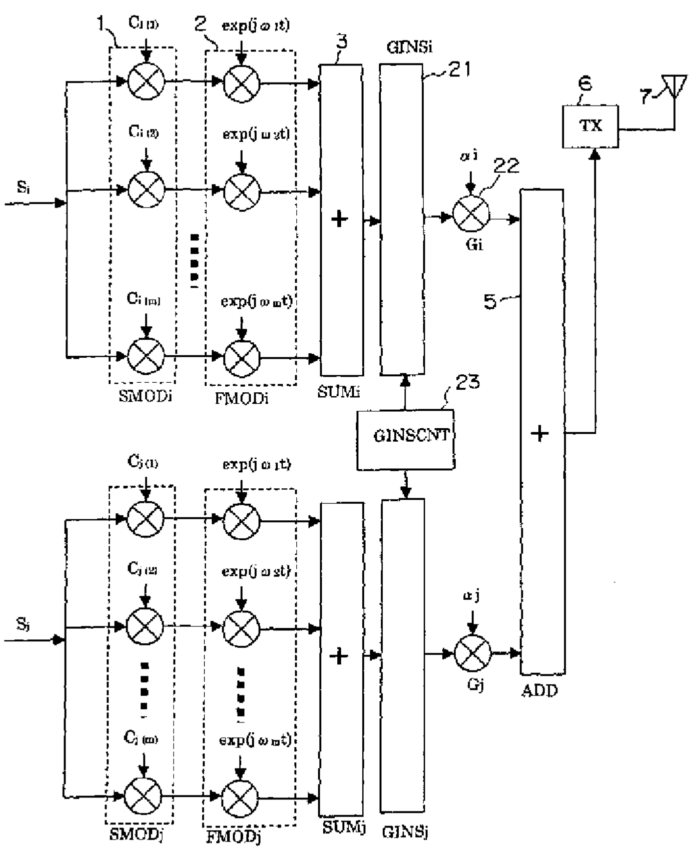
(81) 指定国 (国内): AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO,

[続葉有]

(54) Title: ORTHOGONAL FREQUENCY-DIVISION MULTIPLEX TRANSMISSION METHOD

(54) 発明の名称: 直交周波数分割多重伝送方法



(57) Abstract: A spreading modulator (1) spreads the spectrum of a signal series. A subcarrier modulator (2) modulates the frequencies of a plurality of subcarriers having frequencies different from one another by using the output of the spreading modulator (1). An adder (3) combines the subcarriers modulated. A guard section control unit (23) determines the length of a guard section in accordance with the maximum transmission delay difference of a line between a transmission device and a reception device. A guard section inserter (21) is controlled by the guard section control unit (23) to insert the guard section into the signal series for every symbol period. A gain adjustor (22) multiplies the transmission signal by a gain coefficient corresponding to the guard section inserted.

WO 03/047140 A1

[続葉有]

NZ, PH, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, TZ, UA, UG, US, UZ, VN, YU, ZA, ZW.

(84) 指定国 (広域): ARIPO 特許 (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZM, ZW), ユーラシア特許 (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), ヨーロッパ特許 (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR), OAPI 特許 (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

添付公開書類:
— 国際調査報告書

２文字コード及び他の略語については、定期発行される各*PCT*ガゼットの巻頭に掲載されている「コードと略語のガイダンスノート」を参照。

(57) 要約:

拡散変調器（１）は、信号系列を拡散する。副搬送波変調器（２）は、拡散変調器（１）の出力を用いて互いに周波数の異なる複数の副搬送波を周波数変調する。加算器（３）は、変調された各副搬送波を合成する。ガード区間制御部（２３）は、送信装置と受信装置との間の回線の最大伝送遅延差に応じてガード区間の長さを決定する。ガード区間挿入器（２１）は、シンボル周期ごとに、ガード区間制御部（２３）の制御に従って信号系列にガード区間を挿入する。利得調整器（２２）は、挿入されたガード区間に対応する利得係数を送信信号に乗算する。

明細書

直交周波数分割多重伝送方法

技術分野

本発明は、直交周波数分割多重・符号拡散（ＯＦＤＭ－ＣＤＭ）伝送方式、並びにそのための送信装置（変調装置）及び受信装置（復調装置）に係わり、特に、セルラ電話システムまたは移動体通信システムにおける基地局と移動局との間の通信を実現する装置および方法に係わる。

背景技術

従来より、地上系デジタルテレビ等において、直交周波数分割多重（以下、ＯＦＤＭ：Orthogonal Frequency Division Multiplex）伝送方式が適用されている。ＯＦＤＭ伝送方式では、データは、互いに周波数の異なる複数の副搬送波を利用して伝送される。具体的には、この方式では、互いに直交する多数の副搬送波を送信データで変調し、それらの副搬送波が周波数多重されて伝送される。そして、ＯＦＤＭ伝送方式によれば、高速データの伝送を行う場合においても、各副搬送波ごとの伝送レートを低くできるので、すなわち各副搬送波ごとのシンボル周期を長くできるので、マルチパス干渉の影響が軽減される。なお、ＯＦＤＭ伝送方式につては、例えば、"Overview of Multicarrier CDMA" (Hara et al., IEEE Communication Magazine, Dec. 1997, pp126-133)、あるいは、"WIDEBAND WIRELESS DIGITAL COMMUNICATIONS", A.F.Molisch Prentice Hall PTR, 2001, ISBN:0-13-022333-6)に記載されている。

図１は、ＯＦＤＭ伝送システムにおいて使用される既存の送信装置の構成図である。ここでは、この送信装置は、信号系列Ｓｉ および信号系列Ｓｊ を多重化

して出力するものとする。なお、信号系列Si および信号系列Sj のシンボル周期は「T」であるものとする。また、信号系列Si および信号系列Sj は、例えば、互いに異なる移動機へ送信すべき信号であってもよい。あるいは、信号系列Si内に複数の移動機へ送信すべきデータが時間多重されていてもよい。

信号系列Siの各シンボル情報は、それぞれ拡散変調器1が備えるm個の入力端子に並列に入力される。すなわち、拡散変調器1の各入力端子には、シンボル周期Tとごに、同一のシンボル情報が並列に入力される。そして、拡散変調器1は、入力されたシンボル情報を信号系列Siに対して予め割り当てられている拡散符号Ciを用いて変調し、その結果として得られるmビットの拡散信号を出力する。なお、拡散符号Ci は、「Ci(1)」～「Ci(m)」から構成されており、直交符号列の中の1つの要素であるものとする。

副搬送波変調器2は、互いに異なる角周波数ω1～ωm を持ったm個の副搬送波を生成する。ここで、ω1, ω2, ω3, . . . ωm の角周波数間隔Δωは、シンボル周期Tの逆数により定義される一定の値であり、下記の式により表される。

$$\Delta\omega = 2\pi\Delta f = 2\pi / T$$

また、副搬送波変調器2は、拡散変調器1から出力される拡散信号を用いてm個の副搬送波を変調する。具体的には、例えば、角周波数ω1 を持った副搬送波は、「Ci(1)」が乗算されたシンボル情報により変調され、角周波数ωm を持った副搬送波は、「Ci(m)」が乗算されたシンボル情報により変調される。そして、各副搬送波は、加算器3により合成される。

ガード区間挿入器4は、図2に示すように、シンボル毎に、加算器3から出力される合成信号に対して、予め固定的に決められているガード区間（Guard Interval）を挿入する。ここで、このガード区間は、無線伝送路のマルチパスによる影響を排除するために挿入される。なお、図2では、副搬送波ごとにガード区間が挿入された状態が描かれているが、実際には、これらの副搬送波は合

成されている。

加算器５は、上述のようにして得られる信号系列Ｓｉに対応する合成信号、および同様の処理により得られる信号系列Ｓｊに対応する合成信号を加算する。ここで、信号系列Ｓｉに対応する合成信号および信号系列Ｓｊに対応する合成信号には、それぞれガード区間が挿入されている。そして、加算器５の出力は、送信機６により所定の高周波信号に変換された後、アンテナ７を介して送信される。

図３は、ＯＦＤＭ伝送システムにおいて使用される既存の受信装置の構成図である。ここでは、この受信装置は、図１に示す送信装置により送信された無線信号から信号系列Ｓｉを受信するものとする。なお、図３では、信号を受信するために必要な周波数同期機能、およびタイミング同期機能などは省略されている。

アンテナ１１により受信された信号は、受信機１２によりベースバンド信号Ｓrx に変換された後、副搬送波復調器１３によりｍ個の受信信号列に変換される。続いて、ガード区間削除器１４は、各受信信号列からそれぞれガード区間を削除する。また、拡散復調器１５は、各受信信号系列を逆拡散するために、送信装置において使用された拡散符号と同じ拡散符号Ｃｉ を各受信信号列にそれぞれ乗算する。そして、拡散復調器１５から出力される各信号を加算器１６を用いて加算することにより、信号系列Ｓｉが再生される。

上記構成の送信装置および受信装置の間では、信号系列Ｓｉは、図２に示すように、複数の副搬送波ｆ１〜ｆｍ を利用して伝送される。ここで、信号系列Ｓｉは、「＋１」または「－１」の値を有するシンボル情報から構成されている。即ち、信号系列Ｓｉは、シンボル周期Ｔで「＋１」または「－１」に変化する。また、各副搬送波ｆ１〜ｆｍ を利用して伝送される信号は、それぞれ拡散符号Ｃｉ（Ci(1), Ci(2), . . . Ci(m)）により拡散変調されている。なお、図２において、

「＊」が付されているビットは、信号系列Ｓｉが「－１」であることから拡散変調出力が反転（共役）出力になっていることを示している。

伝送される信号には、上述したように、シンボル毎にガード区間が挿入されている。図２に示す例では、シンボル周期Ｔに対してガード区間Ｔｇが挿入されている。したがって、受信装置では、各副搬送波ごとにそれぞれガード区間Ｔｇを除去することにより得られる区間（区間Ｔｓ）について逆拡散／復調処理が行われる。これにより、受信装置においてマルチパス干渉（遅延波により生ずる干渉）が除去される。

ところで、ガード区間Ｔｇは、マルチパス干渉を除去するために挿入されるので、その長さは、伝送路の最大伝送遅延差よりも長く設定される必要がある。ここで、「最大伝送遅延差」とは、送信装置から受信装置へ複数のパスを介して信号が伝送されるときの最小伝搬時間と最大伝搬時間との差を意味する。例えば、図４において、パス１を介して伝送された信号が最も早く受信装置に到着し、パス３を介して伝送された信号が最も遅く受信装置に到着したとすると、最大伝送遅延差は、パス３による伝搬時間とパス１による伝搬時間との差により表される。

ところが、セルラ通信システムでは、通常、１つの基地局からサービスエリア内の複数の移動機に対して無線信号が送信される。そして、基地局から移動機へ伝送される信号の最大伝送遅延差は、一般に、それらの間の距離が離れるほど大きくなる傾向にある。ここで、サービスエリア内のすべての移動機においてマルチパス干渉を除去しようとすると、基地局から最も遠く離れた位置にいる移動機においてマルチパス干渉を除去できるようにしなければならない。したがって、サービスエリア内のすべての移動機においてマルチパス干渉を除去しようとすると、ガード区間Ｔｇは、基地局から最も遠く離れた位置にいる移動機に信号が伝送されたときの最大伝送遅延差よりも大きくする必要がある。

例えば、図５に示す例では、ガード区間Ｔｇは、基地局から移動機ＭＳ３に信号が伝送されたときの最大伝送遅延差よりも大きくする必要がある。

しかし、このようにしてガード区間の差を決定すると、基地局の近くに位置している移動機（図５では、移動機ＭＳ１）に信号を送信する場合には、ガード区間が必要以上に長くなりすぎる。ここで、ガード区間の信号の電力は、受信装置において信号系列を再生する際に使用されない。このため、上述のようにしてガード区間が決定されると、移動機に信号を送信する際に、無駄な電力が必要となってしまう。この結果、通信システム全体の総伝送容量の低減をまねくことになる。

発明の開示

本発明は、直交周波数分割多重・符号拡散（ＯＦＤＭ－ＣＤＭ）伝送方式を利用した通信システムにおいて、信号の伝送効率を向上させることを目的とする。

本発明の通信システムは、直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する通信システムであって、上記送信装置は、信号系列を用いて複数の副搬送波を変調する変調手段と、上記変調手段の出力にガード区間を挿入する挿入手段と、上記ガード区間が挿入された変調信号を送信する送信手段を有し、上記受信手段は、上記送信装置から送信された変調信号について副搬送波ごとにガード区間の削除処理と復調処理を行い信号系列を再生する復調手段を有し、上記ガード区間の長さは、上記送信装置と上記受信装置との間の通信環境に基づいて決定される。

上記通信システムにおいては、送信装置と受信装置との間の通信環境に基づいてガード区間の長さが決定される。すなわち、ガード区間の長さを、送信装置と受信装置との間の通信環境に応じて必要最小限に短くできる。したがって、

通信効率が向上する。

上記構成において、上記送信装置が、上記ガード区間の長さに応じて上記変調信号を送信する際の送信電力を制御する電力制御手段をさらに有するようにしてもよい。この構成によれば、信号系列を送信する際の送信電力を必要最小限に抑えることができるので、信号間の干渉が低減する。

上記構成において、上記受信装置が、上記送信装置から当該受信装置へ信号が伝送されたときの通信品質をモニタするモニタ手段をさらに有し、上記ガード区間の長さが、予め決められた所定の通信品質が満たされるように決定されるようにしてもよい。この構成によれば、所望の通信品質を満たす範囲内で、必要最小限のガード区間を設定できる。

本発明の他の態様の通信システムは、直交周波数分割多重を利用して送信装置から第１の受信装置を含む複数の受信装置へ信号を伝送する通信システムであって、上記送信装置は、第１の受信装置へ伝送する第１の信号系列および第１の受信装置とは異なる他の受信装置へ伝送する第２の信号系列が多重された信号系列を用いて複数の副搬送波を変調する変調手段と、上記第１の信号系列の変調出力に第１のガード区間を挿入するとともに上記第２の信号系列の変調出力に第２のガード区間を挿入する挿入手段と、上記第１のガード区間と第２のガード区間がそれぞれ挿入された変調信号を送信する送信手段を有し、上記第１の受信装置は、上記第１のガード区間の削除処理と復調処理を行い第１の信号系列を再生する復調手段を有し、上記第１のガード区間の長さは、上記送信装置と上記第１の受信装置との間の通信環境に基づいて決定されると共に、上記第２のガード区間の長さは、上記送信装置と上記他の受信装置との間の通信環境に基づいて決定される。この構成によれば、複数の信号系列を時間多重して送信する際に、各信号系列に対して個々に適切なガード区間を設定できる。

図面の簡単な説明

図１は、ＯＦＤＭ伝送システムにおいて使用される既存の送信装置の構成図である。

図２は、既存のＯＦＤＭ伝送システムにおける伝送信号の例である。

図３は、ＯＦＤＭ伝送システムにおいて使用される既存の受信装置の構成図である。

図４は、マルチパスを説明する図である。

図５は、複数の移動機を収容する基地局を示す図である。

図６は、本発明の実施形態の送信装置の構成図である。

図７は、本発明の実施形態の受信装置の構成図である。

図８および図９は、実施形態のＯＦＤＭ伝送システムにおける伝送信号の例である。

図１０は、ガード区間について説明するための図である。

図１１は、副搬送波変調器により実行される逆フーリエ変換を説明する図である。

図１２は、ガード区間を挿入する処理を説明する図である。

図１３は、ガード区間を挿入する処理を実現する構成の実施例である。

図１４は、受信波からガード区間を削除する処理を実現する構成の実施例である。

図１５は、第１の実施例の送信装置の構成図である。

図１６は、第１の実施例の受信装置の構成図である。

図１７は、第１の実施例の通信システムにおける伝送信号を模式的に示す図である。

図１８は、第２の実施例の送信装置の構成図である。

図１９は、第２の実施例の受信装置の構成図である。

図２０は、第２の実施例の通信システムにおける伝送信号を模式的に示す図である。

図２１は、第３の実施例の送信装置の構成図である。

図２２は、第３の実施例の受信装置の構成図である。

図２３は、図２２に示す遅延差検出部の一例の構成図である。

図２４は、遅延差検出部の動作を説明する図である。

図２５は、最大伝送遅延差を検出する実施例である。

図２６は、第４の実施例の送信装置の構成図である。

図２７は、第４の実施例の受信装置の構成図である。

図２８は、図２７に示す距離推定部の一例の構成図である。

図２９は、第５の実施例の送信装置の構成図である。

図３０は、第５の実施例の受信装置の構成図である。

図３１は、図３０に示すタイミング生成部の一例の構成図である。

図３２は、第６の実施例の送信装置の構成図である。

図３３は、第６の実施例の受信装置の構成図である。

図３４は、図３３に示すタイミング生成部の一例の構成図である。

図３５は、第７の実施例の送信装置の構成図である。

図３６は、第７の実施例の受信装置の構成図である。

図３７は、図３６に示す遅延差検出部の動作を示すフローチャートである。

図３８は、第８の実施例の送信装置の構成図である。

図３９は、第８の実施例の受信装置の構成図である。

図４０は、図３９に示す距離推定部の動作を示すフローチャートである。

発明を実施するための最良の形態

本発明の実施形態について図面を参照しながら説明する。以下では、セルラ

通信システムにおいて直交周波数分割多重・符号拡散（ＯＦＤＭ－ＣＤＭ）伝送方式が利用されるものとする。具体的には、例えば、図５に示す基地局と移動機との間の信号伝送のためにＯＦＤＭ－ＣＤＭが利用されるものとする。

図６は、本発明の実施形態の送信装置の構成図である。なお、この送信装置は、図５においては、例えば、基地局装置に相当する。また、この送信装置は、信号系列Ｓi および信号系列Ｓj を多重化して出力するものとする。ここで、信号系列Ｓi および信号系列Ｓj は、例えば、互いに異なる移動機へ送信すべき信号であってもよい。あるいは、信号系列Ｓi または信号系列Ｓj の中にそれぞれ複数の移動機へ送信すべきデータが時間多重されていてもよい。

この送信装置は、送信すべき信号系列毎に、拡散変調器（ＳＭＯＤ：Spread Modulator）１、副搬送波変調器（ＦＭＯＤ：Frequency Modulator）２、加算器（ＳＵＭ）３、ガード区間挿入器（ＧＩＮＳ：Guard Interval Insert Unit）２１、利得調整器（Ｇ）２２を備える。ここで、拡散変調器１、副搬送波変調器２、加算器３については、図１を参照しながら説明したものを使用することができる。すなわち、拡散変調器１は、ｍ個の入力端子を備えており、それらの入力端子には、シンボル周期Ｔとごに、同一のシンボル情報が並列に入力される。そして、拡散変調器１は、入力されたシンボル情報を信号系列Ｓi に対して予め割り当てられている拡散符号Ｃi を用いて変調し、その結果として得られるｍビットの拡散信号を出力する。なお、拡散符号Ｃi は、「Ｃi(1)」～「Ｃi(m)」から構成されており、直交符号列の中の１つの要素であるものとする。

副搬送波変調器２は、互いに異なる角周波数ω1～ωｍを持ったｍ個の副搬送波を生成する。ここで、ω1, ω2, ω3, . . . ωｍの角周波数間隔Δωは、シンボル周期Ｔの逆数により定義される一定の値であり、下記の式により表される。

$$\Delta\omega = 2\pi\Delta f = 2\pi / T$$

また、副搬送波変調器２は、拡散変調器１から出力される拡散信号を用いて

m個の副搬送波を変調する。具体的には、例えば、角周波数ω1を持った副搬送波は、「Ci(1)」が乗算されたシンボル情報により変調され、角周波数ωm を持った副搬送波は、「Ci(m)」が乗算されたシンボル情報により変調される。なお、副搬送波変調器２の処理は、例えば、逆フーリエ変換演算により実現される。そして、副搬送波変調器２から出力される各副搬送波は、加算器３により合成される。

ガード区間挿入器２１は、シンボル毎に、加算器３から出力される合成信号に対して、ガード区間（Guard Interval）を挿入する。ここで、このガード区間は、無線伝送路のマルチパスによる影響を排除するために挿入される。なお、図１に示した既存の送信装置のガード区間挿入器４は、予め固定的に決められたガード区間を挿入するが、実施形態のガード区間挿入器２１は、送信装置と受信装置との間の通信状態に応じて決められるガード区間を挿入する。なお、ガード区間の長さは、ガード区間制御部（ＧＩＮＳＣＮＴ：Guard Interval Control Unit）２３により、信号系列ごとに決定される。

利得調整器２２は、例えば乗算器であり、ガード区間が挿入された信号に利得係数αを乗算する。これにより、送信すべき信号の振幅または電力が調整される。なお、利得係数αは、基本的に、信号系列ごとに挿入されるガード区間の長さに対応して決定される。

上述のようにして得られる各信号系列ごとの合成信号は、図１に示した既存の送信装置と同様に、加算器（ＡＤＤ）５により加算される。そして、加算器５の出力は、送信機（ＴＸ）６により所定の高周波信号に変換された後、アンテナ７を介して送信される。

このように、実施形態の送信装置では、送信すべき信号系列（Ｓi、Ｓj）ごとに、送信装置と受信装置との間の通信状態に応じて決められるガード区間が挿入される。また、送信すべき信号系列（Ｓi、Ｓj）ごとに、挿入されたガード区

間の長さに対応して送信信号の振幅または電力が調整される。

図７は、本発明の実施形態の受信装置の構成図である。ここでは、この受信装置は、図６に示す送信装置により送信された無線信号から信号系列Ｓｉを受信するものとする。なお、この受信装置は、図５においては、例えば、移動機に相当する。また、図７では、信号を受信するために必要な周波数同期機能、およびタイミング同期機能などは省略されている。

アンテナ１１により受信された信号は、受信機（ＲＸ）１２によりベースバンド信号Ｓrx に変換された後、副搬送波復調器（ＦＤＥＭ：Frequency Demodulator）１３によりｍ個の受信信号列に変換される。ここで、副搬送波復調器１３は、ｍ個の入力端子を備えており、それらの入力端子には、同一のベースバンド信号Ｓrx が並列に入力される。そして、副搬送波復調器１３は、ベースバンド信号Ｓrx に対してそれぞれ角周波数ω1〜ωm を持った周期波を乗算することにより、各副搬送波ごとに信号を復調する。なお、副搬送波復調器１３の処理は、例えば、フーリエ変換演算により実現される。

ガード区間削除器３１は、ガード区間制御部（ＧＣＮＴｉ：Guard Interval Control Unit）３２からの指示に従って、各受信信号列からそれぞれガード区間を削除する。なお、ガード区間制御部３２は、送信装置において信号系列Ｓｉに対して挿入されたガード区間の長さを認識しており、その値をガード区間削除器３１に通知する。したがって、ガード区間削除器３１は、送信装置で挿入されたガード区間を適切に除去することができる。

拡散復調器１５は、各受信信号系列を逆拡散するために、送信装置において使用された拡散符号と同じ拡散符号Ｃｉを各受信信号列にそれぞれ乗算する。そして、加算器１６を用いて拡散復調器１５から出力される各信号を加算することにより、信号系列Ｓｉが再生される。

図８および図９は、実施形態のＯＦＤＭ伝送システムにおける伝送信号の例

である。ここで、図８は、最大伝送遅延差の小さい位置にいる移動機（受信装置）へ送信すべき伝送信号を模式的に示しており、図９は、最大伝送遅延差の大きい位置にいる移動機へ送信すべき伝送信号を模式的に示している。なお、図８に示す伝送信号のシンボル周期が「Ｔ1」であるのに対し、図９に示す伝送信号のシンボル周期が「Ｔ2」であるが、これらの周期は互いに同じであってもよいし、互いに異なっていてもよい。

最大伝送遅延差の小さい位置にいる移動機へ信号を送信する場合は、図８に示すように、各副搬送波ごとに、シンボル周期Ｔ1 に対してガード区間Ｔg1 が挿入される。したがって、信号は、区間Ｔs1 を利用して伝送される。一方、最大伝送遅延差の大きい位置にいる移動機へ信号を送信する場合は、図９に示すように、各副搬送波ごとに、シンボル周期Ｔ2 に対してガード区間Ｔg2 が挿入される。したがって、信号は、区間Ｔs2 を利用して伝送される。そして、このとき、ガード区間Ｔg1 は、ガード区間Ｔg2 よりも短く設定される。すなわち、ガード区間の長さは、送信装置から受信装置へ信号が伝送されたときの最大伝送遅延差が大きくなると、それに応じて長くなる。

また、最大伝送遅延差の小さい位置にいる移動機へ信号を送信する場合は、図８に示すように、信号の送信電力は「Ｐ1」に制御される。一方、最大伝送遅延差の大きい位置にいる移動機へ信号を送信する場合は、図９に示すように、信号の送信電力は「Ｐ2」に制御される。ここで、電力Ｐ2 は、電力Ｐ1 よりも大きい。すなわち、信号の送信電力は、送信装置から受信装置へ信号が伝送されたときの最大伝送遅延差が大きくなると、それに応じて大きくなる。

続いて、ガード区間の挿入／除去について説明する前に、ガード区間そのものについて簡単に説明をする。

図１０は、ガード区間について説明するための図であり、受信装置が受信した信号の波形が模式的に示されている。ここで、実線ａは、受信装置に最初に

到着した信号（基準波）の波形を表し、破線ｂは、受信装置に到着した遅延信号（遅延波）の波形を表している。なお、図１０では、１つの遅延波のみが描かれているが、実際には、通常、２以上の遅延波が存在する。

図１０において、時刻Ｔ1以前は、基準波および遅延波がそれぞれ連続したサイン波なので、受信装置は、それらの合成波から対応するシンボル情報を再生することができる。しかし、シンボル情報が「＋１」から「－１」に変化したとき、あるいは「－１」から「＋１」に変化したときは、そのシンボル情報を伝送する信号の位相が転移する。図１０に示す例では、時刻Ｔ1において基準波の位相が転移しており、時刻Ｔ2において遅延波の位相が転移している。すなわち、この場合、時刻Ｔ1と時刻Ｔ2との間の期間では、基準波は位相転移後の情報を伝送しており、遅延波は位相転移前の情報を伝送していることになる。したがって、この期間は、一方の信号波が他方の信号波に対する干渉波となり、受信波からシンボル情報を適切に再生することができないことがある。

上記干渉による影響は、例えば、図１０に示す例では、受信波から信号を再生する際に、時刻Ｔ1と時刻Ｔ2との間の受信波を使用しないことにより回避される。そして、ＯＦＤＭ通信システムでは、この期間を含む所定の期間をガード区間として定義し、受信装置において信号再生が行われないようにしている。したがって、ガード区間の長さは、最初に到着する信号波と最後に到着する遅延波との遅延差（最大伝送遅延差）よりも大きく設定される必要がある。

ところが、上述したように、最大伝送遅延差は、送信装置と受信装置との間の距離などにより変化する。したがって、実施形態の通信システムでは、ガード区間の長さが最大伝送遅延差に対応して決定されるようになっている。

次に、送信装置においてガード区間を挿入する方法を説明する。ここでは、図６に示す副搬送波変調器２の処理は、逆フーリエ変換演算により実現されるものとする。

図１１は、副搬送波変調器２により実行される逆フーリエ変換を説明する図である。ここでは、シンボル周期を「T」、シンボル周期ごとに挿入されるガード区間を「Tg」、シンボル周期ごとの信号時間を「Ts（＝T－Tg)」とする。

副搬送波変調器２には、上述したように、拡散変調器１から出力されるm個の情報が入力される。ここで、各情報は、それぞれ対応する周波数の副搬送波に割り当てらる。すなわち、副搬送波変調器２は、周波数軸上に配置されたm個の信号を受ける。そして、この周波数軸上のm個の信号は、図１１に示すように、シンボル周期Tごとに実行される逆フーリエ変換により、時間軸上のm個のサンプルから構成される信号系列に変換される。このとき、時間軸上のm個のサンプルは、信号時間Ts内に配置される。

図１２は、ガード区間を挿入する処理を説明する図である。ガード区間挿入器２１は、信号時間Ts内に配置されたm個のサンプルを受け取ると、ガード区間Tgに相当する個数のサンプル成分を信号時間Tsの末尾から抽出し、それらを信号時間Tsの直前に複写する。図１２に示す例では、ガード区間Tgが３サンプリング時間に相当し、m個のサンプル「１」～「m」のうちから、「m－２」「m－１」「m」が抽出されて信号時間Ts の直前に複写されている。そして、この複写により、シンボル時間T（＝Tg＋Ts）の時間軸上の信号系列が作成される。

図１３は、ガード区間を挿入する処理を実現する構成の実施例である。上述したように、副搬送波変調器２は、逆フーリエ変換器によって実現され、シンボル周期ごとに、周波数軸上のm個の信号を時間軸上のm個のサンプル（t1～tm）に変換する。そして、ガード区間挿入器２１は、まず、ガード区間Tgにおいて、「tm-2」「tm-1」「tm」を順番に読み出して出力し、それに続く信号時間Ts において「t1」～「tm」を順番に読み出して出力していく。これにより、ガード区間が挿入された信号系列が作成される。

上記構成においてガード区間の長さは、「信号時間Ts の前に出力するサンプルの数」を変えることにより制御される。この場合、ガード区間Tg、信号時間Ts、および逆フーリエ変換の周期（すなわち、シンボル周期T）が所定の関係（T＝Tg＋Ts）を満たすように、サンプル値の読出し間隔が決定される。

一例を示す。ここでは、シンボル周期＝T、ガード区間Tg＝０．２T、信号時間Ts＝０．８T、副搬送波の多重数m＝１０００であるものとする。この場合、ガード区間挿入器２１には、シンボル周期ごとに、時間軸上の１０００個のサンプル（t 1～t 1000）が入力される。そして、まず、２５０（＝１０００×０．２÷０．８）個のサンプル（t 751～t 1000）を読み出して出力する。続いて、上記１０００個のサンプル（t 1～t 1000）を読み出して出力する。このとき、サンプル値の読出し間隔は、「T／１２５０」である。また、ガード区間Tg＝０．１T、信号時間Ts＝０．９T、副搬送波の多重数m＝１０００であるものとすると、ガード区間挿入器２１は、まず、１１１（＝１０００×０．１÷０．９）個のサンプル（t 890～t 1000）を読み出して出力し、それ続いて、上記１０００個のサンプル（t 1～t 1000）を読み出して出力する。このとき、サンプル値の読出し間隔は、「T／１１１１」である。

なお、実施形態では、複数の副搬送波が合成された後にガード区間が挿入されているが、原理的には、副搬送波ごとにガード区間を挿入することも可能である。

図１４は、受信装置において受信波からガード期間を削除する処理を実現する構成の実施例である。ここでは、図１１～図１３に示すようにして作成された信号列（t m-2, t m-1, t m, t 1, t 2, t 3,... t m）が受信されるものとする。なお、図７に示す受信装置では、副搬送波変調を行った後にガード区間が削除されるように描かれているが、図１４に示す構成では、これらの処理は一体的に実行される。

ガード区間削除器３１は、スイッチ４１およびシフトレジスタ４２を備える。そして、信号系列（ｔm-2，ｔm-1，ｔm，ｔ1，ｔ2，ｔ3，．．．ｔm）を受信すると、スイッチ４１を適切にＯＮ／ＯＦＦ制御することにより、ガード区間に配置されている所定数のサンプル値（ここでは、ｔm-2，ｔm-1，ｔm）を廃棄し、後続のm個のサンプル値（ｔ1～ｔm）をシフトレジスタ４２に送る。ここで、ガード区間削除器３１は、送信装置において挿入されたガード区間の長さ（あるいは、ガード区間内のサンプル数）を認識しており、それに基づいてスイッチ４１のＯＮ／ＯＦＦ状態を制御する。一方、副搬送波復調器１３として動作するフーリエ変換器は、シフトレジスタ４２にm個のサンプル値が蓄積されると、それらのサンプル値についてフーリエ変換を行うことにより、副搬送波ごとの信号ｆ1～ｆm を得る。なお、この処理は、シンボル周期Ｔごとに繰り返し実行される。

このように、実施形態のセルラ通信システムでは、送信装置（基地局）から受信装置（移動機）へ信号を送信する際、それらの間の最大伝送遅延差に基づいて、ガード区間の長さ、および送信電力が決定される。ここで、送信装置と受信装置との間の距離が短い場合は、最大伝送遅延差が小さくなり、ガード区間が短くなる。そして、ガード区間が短くなると、それに応じて受信装置において信号再生に寄与する信号時間が長くなるので、送信電力を低くすることができる。したがって、システム全体として干渉電力が減少し、伝送容量が増加することになる。

次に、上述の送信装置および受信装置の実施例を説明する。

第1の実施例：

図１５および図１６は、第１の実施例の送信装置および受信装置の構成図である。これらの装置の基本構成は、それぞれ、図６に示した送信装置および図７に示した受信装置と同じである。ただし、第１の実施形態の送信装置は、時

間多重された複数の信号系列を１つのＯＦＤＭ－ＣＤＭユニット（拡散変調器１、副搬送波変調器２、加算器３、ガード区間挿入器２１）により一括して変調することができる。

すなわち、信号系列Ｓi1 および信号系列Ｓi2 は、図１７に示すように、時間多重化部（TDMi）５１により多重化される。ここでは、これらの信号系列は、互いに異なる最大伝送遅延差を有する回線を介して伝送されるものとする。そして、この信号系列は、拡散変調器１および副搬送波変調器２により変調された後、ガード区間挿入器２１に与えられる。

ガード区間挿入器２１は、入力される信号系列に対して、対応する最大伝送遅延差よりも広いガード区間を挿入する。ここで、各信号系列に対するガード区間は、ガード区間制御部２３により設定される。また、利得調整器２２は、挿入されたガード区間に応じて決まる利得係数αを送信信号に乗算する。具体的には、図１７に示す例では、信号系列Ｓi1 が入力されている期間は、シンボル周期ごとにガード区間Ｔg1 が挿入され、信号の送信電力が「Ｐ1」になるように利得係数$\alpha i(t)$が制御される。一方、信号系列Ｓi2 が入力されている期間は、シンボル周期ごとにガード区間Ｔg2 が挿入され、信号の送信電力が「Ｐ2」になるように利得係数$\alpha i(t)$が制御される。

そして、上述のようにして変調された信号は、他の系の信号と合成された後、アンテナ７を介して送信される。

受信装置の基本的な動作は、図７を参照しながら説明した通りである。ただし、この受信装置は、自分宛ての信号のみを再生する。例えば、信号系列Ｓi1 および信号系列Ｓi2 が時間多重された信号から信号系列Ｓi1 を再生する場合には、ガード区間制御部３２は、信号系列Ｓi1 を受信している期間に、ガード区間Tg1 を削除するようにガード期間削除器３１に対して指示を与える。そして、ガード区間削除器３１は、その指示に従って信号系列Ｓi1 のシンボル周期ごと

にガード区間を削除する。このとき、信号系列Si2 を受信している期間は、ガード区間は削除される必要はない。

ガード区間削除器３１の出力は、拡散復調器１５により逆拡散復調される。このとき、拡散復調器１５は、ガード区間Tg1 が削除された信号時間Ts1 について逆拡散復調を行う。そして、分離部（DML）５２は、復調された信号から、信号系列Si1 に対応する時間スロットにおいてデータを出力する。

このように、第１の実施例の通信システムでは、時間多重された複数の信号系列を１つのOFDM－CDMユニット（拡散変調器１、副搬送波変調器２、加算器３、ガード区間挿入器２１）により一括して変調できる。

第２の実施例：

第２の実施例の通信システムは、第１の実施例の通信システムの変形例である。すなわち、第１の実施例のシステムでは、時間多重された信号系列Si1 および信号系列Si2 がOFDM－CDMを利用して伝送される。ここで、信号系列Si1 および信号系列Si2 は、基本的に、それぞれ対応する移動機に送信されることを想定している。これに対して、第２の実施例のシステムでは、時間多重された報知情報Bi および信号系列Si1 がOFDM－CDMを利用して伝送される。ここで、信号系列Si1 は、所定の１または複数の受信装置に対して送信されるが、報知情報Bi は、サービスエリア内のすべての受信装置（移動機）に対して送信される。したがって、この報知情報Bi は、サービスエリア内の最も遠くに位置する受信装置（すなわち、最大伝送遅延差が最も大きくなる受信装置）に適切に伝送されるようなガード区間が設定され、且つ、送信電力が決定される必要がある。

図１８および図１９は、第２の実施例の送信装置および受信装置の構成図である。これらの装置の基本構成は、それぞれ、図１５に示した送信装置および図１６に示した受信装置と同じである。

第２の実施例では、ガード区間挿入器２１は、図２０に示すように、ガード区間制御部２３からの指示に従って、報知情報Ｂｉが入力されている期間は、シンボル周期ごとにガード区間Ｔg1を挿入し、信号系列Ｓi1が入力されている期間は、シンボル周期ごとにガード区間Ｔg2 を挿入する。ここで、報知情報Ｂｉに対して挿入されるガード区間Ｔg1 は、サービスエリア内において生じる最も大きな最大伝送遅延差よりも長くなるように設定される。例えば、図５において、基地局から移動機ＭＳ１～ＭＳ３へ報知情報を送信する際、基地局から移動機ＭＳ３への回線の最大伝送遅延差が最も大きかったとすると、ガード区間Ｔg1 の長さは、その最大伝送遅延差よりも長くなるように設定される。一方、信号系列Ｓi1 に対して挿入されるガード区間Ｔg2 は、対応する受信装置への回線の最大伝送遅延差よりも長くなるように設定される。例えば、図５において、基地局から移動機ＭＳ１信号系列Ｓi1 を送信する際には、ガード区間Ｔg2 の長さは、基地局から移動機ＭＳ１への回線の最大伝送遅延差よりも長くなるように設定される。

また、利得調整器２２は、ガード区間挿入器２１により挿入されたガード区間に応じた利得係数αを送信信号に乗算する。具体的には、図２０に示す例では、利得係数αi(t)は、報知情報Ｂｉを伝送するための信号の送信電力が「Ｐ1」となり、信号系列Ｓi1 を伝送するための信号の送信電力が「Ｐ2」になるように制御される。したがって、このように制御される利得係数αを送信信号に乗算することにより、報知情報Ｂｉはサービスエリア内のすべての受信装置に伝送されるように大きな送信電力で送信され、信号系列Ｓi1 は対応する受信装置に伝送される範囲で必要最小限の送信電力で送信される。

受信装置では、ガード区間制御部３２は、報知情報Ｂｉを受信している期間はガード区間Ｔg1 を指示し、信号系列Ｓi1 を受信している期間はガード区間Ｔg2 を指示する。そして、ガード区間削除器３１は、ガード区間制御部３２からの

指示に従って受信信号からガード区間を削除する。さらに、ガード区間が削除された信号は、拡散復調器１５により逆拡散された後、分離部５２により報知情報Ｂｉおよび信号系列Ｓｉ1に分離される。

なお、報知情報Ｂｉに対して挿入されるガード区間Ｔg1の長さは、例えば、以下のようにして決定される。

（１）通信エリアの大きさに基づいて決定する。すなわち、送信装置がカバーする通信エリアの大きさに基づいて、報知情報Ｂｉが最も遅延して到着する受信装置までの遅延時間を推定し、その遅延時間に従ってガード区間Ｔg1の長さを決定する。

（２）報知情報Ｂｉ を送信する際の送信装置の送信電力に基づいて決定する。すなわち、報知情報Ｂｉ の送信電力により、その報知情報Ｂｉ を複数の受信装置に送信する際の伝送遅延時間の最大値を推定し、その遅延時間に従ってガード区間Ｔg1の長さを決定する。

（３）通信エリア内に存在する複数の受信装置との間の通信環境に基づいて決定する。すなわち、送信装置がカバーする通信エリア内に存在する複数の受信装置との間の通信環境をそれぞれ求め、これに基づいてガード区間Ｔg1の長さを決定する。具体的には、通信環境が最も厳しい受信装置に合わせてガード区間Ｔg1の長さを決定する。

（４）通信エリア内の最大遅延時間に基づいて決定する。すなわち、送信装置から通信エリア内に存在する複数の受信装置へ報知情報Ｂｉ を送信したときの遅延時間を受信装置ごとに測定し、それらのうちの最大遅延時間に基づいてガード区間Ｔg1の長さを決定する。

第３の実施例：

第３の実施例の通信システムでは、送信装置から受信装置へ信号が伝送されたときの最大伝送遅延差を検出し、その検出結果に基づいてガード区間および

送信電力が決定される。したがって、第３の実施例における送信装置および受信装置は、そのための機能を備えている。

図２１は、第３の実施例の送信装置の構成図である。この送信装置は、対応する受信装置において検出された最大伝送遅延差を表す最大伝送遅延差情報（τ）を受け取り、その情報に基づいてガード区間および送信電力を決定する機能を備えている。即ち、ガード区間制御部（ＧＩＮＳＣＮＴ）６１は、対応する受信装置において検出された最大伝送遅延差に基づいて、挿入すべきガード区間の長さを決定する。具体的には、ガード区間制御部６１ｉは、信号系列Ｓi1 および／または信号系列Ｓi2 を受信する受信装置から送られてくる最大伝送遅延差情報（τi）に基づいて、信号系列Ｓi1 および／または信号系列Ｓi2 を伝送するための信号に挿入すべきガード区間を決定する。また、電力制御部（ＰＣＮＴ）６２は、対応する受信装置において検出された最大伝送遅延差に基づいて、利得係数αを決定する。具体的には、電力制御部６２ｉは、信号系列Ｓi1 および／または信号系列Ｓi2 を受信する受信装置から送られてくる最大伝送遅延差情報（τi）に基づいて、信号系列Ｓi1 および／または信号系列Ｓi2 を伝送するための信号に乗算すべき利得係数αを決定する。

そして、ガード区間挿入器２１は、シンボル周期ごとに、送信信号に対してガード区間制御部６１により決定されたガード区間を挿入する。また、利得調整器２２は、電力制御部６２により決定された利得係数αを送信信号に乗算することにより、ガード区間の長さに対応する送信電力を実現する。

図２２は、第３の実施例の受信装置の構成図である。この受信装置は、送信装置から送られてきた信号の最大伝送遅延差を検出する機能を備えている。すなわち、遅延差検出部（ＤＭＥＳ）６３は、受信したベースバンド信号Ｓrx から最大伝送遅延差を検出し、その検出結果を表す最大伝送遅延情報をガード区間制御部６４および対応する送信装置に通知する。ガード区間制御部６４は、

遅延差検出部６３からの通知に従ってガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図２３は、図２２に示す遅延検出部６３の一例の構成図である。遅延差検出部６３は、ベースバンド信号Ｓrx を時間Ｔs だけ遅延させる遅延回路７１、乗算器７２ａおよび積分器７２ｂから構成される相関検出回路７２、相関検出回路７２により検出された相関値と予め決められている所定のしきい値とを比較する比較回路７３、および比較回路７３による比較結果に基づいて最大伝送遅延差を検出する検出回路７４を含む。ここで、乗算器７２ａは、ベースバンド信号Ｓrx にその遅延信号を乗算し、積分器７２ｂは、乗算器７２ａの出力を積分する。以下、図２４を参照しながら遅延差検出部６３の動作を説明する。

相関検出回路７２には、ベースバンド信号Ｓrx およびそのベースバンド信号Ｓrx を時間Ｔs だけ遅延させた信号（遅延信号）が入力される。ここで、各シンボル周期内のガード区間Ｔg には、図１１～図１３を参照しながら説明したように、信号時間Ｔs の最後尾部分のサンプル値が複写されている。このため、ベースバンド信号Ｓrx とその遅延信号との間では、ベースバンド信号Ｓrx の最後尾部分と遅延信号のガード区間とが重なったときに相関（自己相関）が高くなる。ただし、送信装置と受信装置との間に伝送遅延の異なる複数のパスが存在する場合には、各パスを介して信号を受信するごとに相関値のピークが発生する。したがって、比較回路７３を用いて上記相関値と予め設定されているしきい値とを比較すれば、各パスを介して信号を受信したタイミングをそれぞれ検出できる。よって、最初に信号を受信したタイミングと、最後に信号を受信したタイミングとの時間差を測定することにより、最大伝送遅延差が検出される。例えば、図４に示す通信環境においては、図２５に示すようにして最大伝送遅延差が検出される。

このように、第３の実施例では、送信装置と受信装置との間の回線の最大伝送遅延差が測定され、その結果に基づいてガード区間が挿入／削除されるので、ガード区間の幅を動的に変化させることが可能である。また、上記最大伝送遅延差の測定結果に従って送信信号の利得係数が決定されるので、常に、送信電力を必要最小限に抑えられる。

第４の実施例：

第４の実施例の通信システムでは、送信装置と受信装置との間の伝送距離を推定し、その推定結果に基づいてガード区間および送信電力が決定される。したがって、第４の実施例における送信装置および受信装置は、そのための機能を備えている。

図２６は、第４の実施例の送信装置の構成図である。この送信装置は、対応する受信装置との間の伝送距離の推定値を表す伝送距離情報（Ｌ）を受け取り、その情報に基づいてガード区間および送信電力を決定する機能を備えている。即ち、ガード区間制御部（ＧＩＮＳＣＮＴ）８１は、送信装置と受信装置との間の伝送距離に基づいて、挿入すべきガード区間の長さを決定する。具体的には、ガード区間制御部８１ｉは、信号系列Si1および／または信号系列Si2を受信する受信装置から送られてくる伝送距離情報（Li）に基づいて、信号系列Si1および／または信号系列Si2を伝送するための信号に挿入すべきガード区間を決定する。また、電力制御部（ＰＣＮＴ）８２は、上記伝送距離に基づいて、利得係数αを決定する。具体的には、電力調整部８２ｉは、信号系列Si1および／または信号系列Si2を受信する受信装置から送られてくる伝送距離情報（Li）に基づいて、信号系列Si1および／または信号系列Si2を伝送するための信号に乗算すべき利得係数αを決定する。

そして、ガード区間挿入器２１は、シンボル周期ごとに、送信信号に対してガード区間制御部８１により決定されたガード区間を挿入する。また、利得調

整器２２は、電力制御部８２により決定された利得係数αを送信信号に乗算することにより、ガード区間の長さに対応する送信電力を実現する。

図２７は、第４の実施例の受信装置の構成図である。この受信装置は、送信装置と当該受信装置との間の伝送距離を推定する機能を備えている。すなわち、距離推定部（ＬＭＥＳ）８３は、受信したベースバンド信号Ｓrx に基づいて送信装置と当該受信装置との間の伝送距離を推定し、その推定結果を表す伝送距離情報Ｌをガード区間制御部８４および対応する送信装置に通知する。ガード区間制御部８４は、距離推定部８３からの通知に従ってガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図２８は、図２７に示す距離推定部８３の一例の構成図である。距離推定部８３は、第３の実施例において説明した遅延差検出部６３および変換テーブル８５から構成される。

送信装置と受信装置との間の伝送距離は、その間の回線の最大伝送遅延差と相関があり、伝送距離が長くなるほど最大伝送遅延差も大きくなることが知られている。したがって、これらの間の関係を実験またはシミュレーション等により予め求めておけば、最大伝送遅延差を検出することによって伝送距離を推定することができる。このため、距離推定部８３の変換テーブル８５には、伝送距離と最大伝送遅延差との関係を表す情報が格納されている。そして、遅延差検出部６３により検出された最大伝送遅延差をキーとしてその変換テーブル８５を検索することにより、送信装置と受信装置との間の伝送距離が推定される。

第５の実施例：

第５の実施例の通信システムでは、第４の実施例と同様に、送信装置と受信装置との間の伝送距離を推定し、その推定結果に基づいてガード区間および送

信電力が決定される。ただし、第５の実施例における推定方法は、第４の実施例のそれと異なっている。

図２９は、第５の実施例の送信装置の構成図である。この送信装置は、対応する受信装置からタイミング情報（T）を受け取ってそれに基づいて送信装置と受信装置との間の伝送距離を推定する機能、およびその伝送距離の推定値に基づいてガード区間および送信電力を決定する機能を備えている。

ガード区間制御部（ＧＩＮＳＣＮＴ）９１または電力制御部（ＰＣＮＴ）９２は、対応する受信装置から送られてくるタイミング信号Ｔに基づいて、当該送信装置と対応する受信装置との間の距離を推定する。すなわち、第５の実施例では、送信装置から信号が送信され、その信号が対応する受信装置により検出され、さらにその受信装置において上記信号が検出された旨が送信装置に通知される。ここで、上記信号が上記受信装置において検出されたタイミングは、タイミング情報Ｔを用いて送信装置に通知される。したがって、ガード区間制御部９１または電力制御部９２は、信号を送信したときから、対応する受信装置からタイミング情報Ｔを受信するまでの時間をモニタすることにより、送信装置と受信装置との間の伝送時間および伝送距離を推定できる。そして、上記伝送距離の推定値は、伝送距離情報Ｌを利用して対応する受信装置に送られる。

なお、ガード区間制御部９１は、伝送距離の推定値に基づいてガード区間の長さを決定する。また、電力制御部９２は、伝送距離の推定値に基づいて利得係数αを決定する。これらの処理は、基本的に、第４の実施例と同じである。

図３０は、第５の実施例の受信装置の構成図である。この受信装置は、送信装置から送出された信号の受信タイミングを検出する機能を備えている。すなわち、タイミング生成部（ＴＧＥＮ）９３は、受信したベースバンド信号Ｓrxを基準として受信タイミングを検出し、タイミング信号Ｔを生成する。そして、生成したタイミング信号Ｔは、送信装置へ送られる。また、ガード区間制御部

（ＧＣＮＴ）９４は、送信装置から送られてくる伝送距離情報Ｌに基づいてガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図３１は、図３０に示すタイミング生成部９３の一例の構成図である。タイミング生成部９３は、第３の実施例において説明した遅延回路７１、相関検出回路７２、および最大値判定回路９５を含む。

上述したように、受信信号とその遅延信号との自己相関をモニタした場合、ガード区間を受信している期間の相関値が高くなる。したがって、その相関値をモニタすることにより、ガード区間の位置を検出できる。具体的には、最大値判定部９５を用いてシンボル周期ごとに上記相関値の最大値を検出することにより、ガード区間のタイミング（または、ガード区間の直後に相当するタイミング）を検出できる。そして、タイミング生成部９３は、検出したタイミングを表すタイミング情報Ｔを生成し、それを送信装置へ送る。

第６の実施例：

第６の実施例の通信システムでは、第４または第５の実施例と同様に、送信装置と受信装置との間の伝送距離を推定し、その推定結果に基づいてガード区間および送信電力が決定される。ただし、第６の実施例における推定方法は、第４または第５の実施例のそれと異なっている。

第６の実施例の通信システムでは、信号系列Ｓi1 および信号系列Ｓi2 を送信する際に、それらの系列にそれぞれ既知情報ＳＷが時間多重される。一方、受信装置は、受信信号に中に含まれている既知情報ＳＷを検出すると、その検出タイミングを送信装置に通知する。そして、送信装置は、既知情報ＳＷを送信したタイミングおよび対応する受信装置から送られてくるタイミング情報に基づいて、当該送信装置と受信装置との間の信号の伝送時間を検出し、その伝送時間から伝送距離を推定する。

図３２は、第６の実施例の送信装置の構成図である。この送信装置は、送信信号系列に既知情報ＳＷを多重化する機能、対応する受信装置からタイミング情報（Ｔ）を受け取ってそれに基づいて送信装置と受信装置との間の伝送距離を推定する機能、およびその伝送距離の推定値に基づいてガード区間および送信電力を決定する機能を備えている。

時間多重化部（TDM）５１は、信号系列Si1、Si2 を送信する際に、それらの系列にそれぞれ既知情報ＳＷを多重する。ここで、既知情報ＳＷは、特に限定されるものではないが、対応する受信装置がそのデータパターンを認識している必要がある。

ガード区間制御部（ＧＩＮＳＣＮＴ）１０１または電力制御部（ＰＣＮＴ）１０２は、対応する受信装置から送られてくるタイミング信号Ｔに基づいて、当該送信装置と対応する受信装置との間の距離を推定する。そして、この伝送距離の推定値は、伝送距離情報Ｌを利用して対応する受信装置に送られる。なお、伝送距離を推定する方法については後述する。

なお、ガード区間制御部１０１は、伝送距離の推定値に基づいてガード区間の長さを決定する。また、電力制御部１０２は、伝送距離の推定値に基づいて利得係数αを決定する。これらの処理は、基本的に、第４または第５の実施例と同じである。

図３３は、第６の実施例の受信装置の構成図である。この受信装置は、受信波から既知情報ＳＷを分離して出力する機能、および既知情報を受信した旨を送信装置に通知する機能を備えている。すなわち、タイミング生成部（ＴＧＥＮ）１０３は、分離部（ＤＭＬ）５２から出力された既知情報ＳＷを検出すると、その検出タイミングから所定時間経過後にタイミング信号Ｔを生成して送信装置へ送出する。また、ガード区間制御部（ＧＣＮＴ）１０４は、送信装置から送られてくる伝送距離情報Ｌに基づいてガード区間を決定し、それをガー

ド区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図３４は、図３３に示すタイミング生成部１０３の一例の構成図である。タイミング生成部１０３には、当該受信装置により復調された信号列が入力される。ここで、この信号列は、送信装置において挿入された既知情報ＳＷを含んでいる。そして、この信号列は、既知情報ＳＷのワード長と等しい長さのシフトレジスタ１０５に順番に入力されていく。論理反転回路１０６、加算回路１０７、および比較回路１０８は、シフトレジスタ１０５に新たなデータの書き込まれるごとに、保持されているデータが既知情報ＳＷと一致するか否かを調べる。なお、論理反転回路１０６は、既知情報ＳＷのワードパターンに対応して設けられている。また、加算回路１０７は、シフトレジスタ１０５に保持されている各エレメントの値またはシフトレジスタ１０５に保持されている各エレメントの値の論理反転値を加算する。そして、比較回路１０８は、加算回路１０７による加算結果と予め設定されている閾値とを比較し、加算結果の方が大きかったときにタイミング信号Ｔを出力する。

このように、第６の実施例の通信システムでは、送信装置から受信装置へ既知情報ＳＷが送信され、その既知情報ＳＷを検出した旨が受信装置から送信装置へ通知される。したがって、送信装置から受信装置へ信号が伝送される際の伝送時間を「Ｔ1」、受信装置が既知情報ＳＷを検出してからタイミング情報を送信するまでの時間を「Ｔd」、受信装置から送信装置へタイミング情報が伝送される際の伝送時間を「Ｔ2」、既知情報ＳＷを送信してからタイミング情報を受信するまでの時間を「Ｔ0」とすると、下記の式が成立する。なお、「Ｔ2」は「Ｔ1」に比例するものとし、その比例定数を「β」とする。

$$
\begin{aligned}
T1 &= T0 - Td - T2 \\
&= T0 - Td - \beta \cdot T1
\end{aligned}
$$

$\therefore\ T1 = (T0 - Td) / (1 + \beta)$

ここで、送信装置と受信装置との間の伝送距離は、送信装置から受信装置へ信号が伝送される際の伝送時間（T1）に比例する。また、受信装置が既知情報ＳＷを検出してからタイミング情報を送信するまでの時間（Td）は既知である。したがって、送信装置は、既知情報ＳＷを送信してからタイミング情報を受信するまでの時間（T0）を測定することにより、送信装置と受信装置との間の伝送距離を推定できる。なお、この実施例では、ガード区間制御部１０１または電力制御部１０２がその伝送距離を推定する。

第７の実施例：

第７の実施例の通信システムでは、ガード区間の長さを変えながら伝送エラー率が測定され、所定の伝送品質が確保されるようにガード区間の長さ（および、送信電力）が決定される。したがって、第７の実施例における送信装置および受信装置は、そのための機能を備えている。

図３５は、第７の実施例の送信装置の構成図である。この送信装置は、既知パターンデータ（ＰＬj）を変調して送信する機能、および対応する受信装置から最大伝送遅延差情報（τ）を受け取ってそれに基づいてガード区間および送信電力を決定する機能を備えている。

既知パターンデータ（ＰＬj）は、拡散変調器１により拡散された後、副搬送波変調器２により変調される。ここで、既知パターンデータ（ＰＬj）は、特に限定されるものではないが、各受信装置により認識されているものとする。また、拡散変調器１は、既知パターンデータ（ＰＬj）に対応する拡散符号Ｃ（ＰＬj）により拡散される。

ガード区間挿入器（ＧＩＮＳj）２１は、シンボル周期ごとに、既知パターンデータ（ＰＬj）を伝送するための信号系列に比較的長いガード区間を挿入する。ここで、このガード区間は、例えば、サービスエリア内の最も遠い位置にいる

移動機（受信装置）へ信号を送信する場合を想定して決定されるようにしてもよい。また、利得調整器（Gj）２２は、ガード区間が挿入された信号系列が十分に大きな送信電力で送信されるように適切な利得係数αjを乗算する。ここで、この利得係数αjは、例えば、サービスエリア内の最も遠い位置にいる移動機（受信装置）へ信号を送信する場合を想定して決定されるようにしてもよい。そして、既知パターンデータ（ＰＬj）は、信号系列Ｓi1、Ｓi2 と合成されて送信される。

ガード区間制御部（ＧＩＮＳＣＮＴ）６１および電力制御部（ＰＣＮＴ）６２の動作は、第３の実施例において説明した通りである。すなわち、ガード区間制御部６１は、対応する受信装置から送られてくる最大伝送遅延差情報に基づいて、挿入すべきガード区間の長さを決定する。また、電力制御部６２は、対応する受信装置から送られてくる最大伝送遅延差情報に基づいて、利得係数αを決定する。

図３６は、第７の実施例の受信装置の構成図である。この受信装置は、既知パターンデータ（ＰＬj）を抽出してその伝送エラーを測定する機能、および伝送エラー率に基づいて最大伝送遅延差情報を生成する機能を備えている。

受信波は、復調回路により復調される。このとき、拡散復調器（ＳＤＥＭ）１５において、信号系列Ｓi1 を復調するときは拡散符号Ｃi が使用され、既知パターンデータ（ＰＬj）を復調するときには拡散符号Ｃ（ＰＬj）が使用される。そして、分離部５２は、再生された信号列を、信号系列Ｓi1 および既知パターンデータ（ＰＬj）に分離する。

遅延差検出部（ＤＭＥＳ）１１１は、再生された既知パターンデータ（ＰＬj）の伝送エラー率を測定し、その伝送エラー率に基づいて最大伝送遅延差情報τを生成する。この最大伝送遅延差情報τは、ガード区間制御部（ＧＣＮＴ）１１２に与えられると共に、送信装置に送られる。そして、ガード区間制御部１

１２は、その最大伝送遅延差情報τに基づいてガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図３７は、図３６に示す遅延差検出部１１１の動作を示すフローチャートである。ここでは、予め複数のガード区間長データτ0～τnが用意されているものとする。また、ガード区間長データτ0～τn の中で、「τ0」が最小であり、「τn」が最大であるものとする。なお、このフローチャートの処理は、たとえば、既知パターンデータ（ＰＬj）を受信するごとに実行される。

ステップＳ１では、拡散復調器１５に拡散符号Ｃ（ＰＬj）を設定する。ここで、この拡散符号Ｃ（ＰＬj）は、送信装置において既知パターンデータ（ＰＬj）を拡散する際に使用されてものである。これにより、以降、受信信号が逆拡散されると、既知パターンデータ（ＰＬj）が再生されることになる。ステップＳ２では、ガード区間長データを指定するための変数を初期化する。すなわち、「ｉ＝０」が設定される。

ステップＳ３では、ガード区間制御部１１２にガード区間長データτｉを設定する。ただし、この時点では、「ｉ＝０」であるので、ガード区間制御部１２３には「ガード区間長データτ0」が設定されることになる。ここで、「ガード区間長データτ0」は、予め用意されている候補データの中で最も短い値を持っている。また、このとき、分離部５２は、再生された既知パターンデータ（ＰＬj）が遅延差検出部１１１に導かれるように出力する。

ステップＳ４では、再生された既知パターンデータ（ＰＬj）の誤り率（誤りビット数）を調べる。そして、この誤り率が予め設定されているしきい値よりも高かった場合には、十分な通信品質が得られていないものとみなし、ステップＳ５へ進む。ステップＳ５では、変数ｉをインクリメントできるか否かが調べられる。そして、可能であれば、ステップＳ６において変数ｉがインクリメ

ントされた後、ステップＳ３に戻る。

このように、ステップＳ３～Ｓ６では、ガード区間制御部１１２に設定すべきガード区間長を少しずつ長くしていきながら、それぞれについて既知パターンデータ（ＰＬj）の誤り率が測定される。そして、既知パターンデータ（ＰＬj）の誤り率がしきい値以下になった時点で、ステップＳ７へ進む。したがって、上記処理により、所望の通信品質が得られる範囲内で、できるかぎり短いガード区間長が決定される。なお、この時点で、ガード区間制御部１１２には、最適なガード区間が設定されていることになる。

ステップＳ７では、拡散復調器１５に拡散符号Ｃｉを設定する。ここで、拡散符号Ｃｉは、送信装置において信号系列Ｓi1、Ｓi2を拡散する際に使用されたものである。したがって、以降、拡散復調器１５は、受信信号から信号系列Ｓi1を復調できるようになる。ステップＳ８では、ステップＳ３～Ｓ６において決定されたガード区間長を送信装置に通知する。

このように、第７の実施例では、伝送エラー率を測定しながら所定の通信品質が確保されるようにガード区間の長さ（および、送信電力）が決定される。したがって、必要最小限のガード区間および送信電力で所望の通信品質が確保される。

第８の実施例：

第８の実施例の通信システムは、第７の実施例の通信システムの変形例である。すなわち、第７の実施例では、受信装置に設定すべきガード区間長が決定され、その値が送信装置に通知される構成であった。これに対して、第８の実施例では、受信装置に設定すべきガード区間長に基づいて送信装置と受信装置との間の伝送距離が推定され、その推定結果が送信装置に通知される。

図３８は、第８の実施例の送信装置の構成図である。この送信装置は、基本的には、図３５に示した第７の実施例の送信装置と同じである。ただし、第８

の実施例の送信装置は、図３５に示したガード区間制御部（ＧＩＮＳＣＮＴ）６１および電力制御部（ＰＣＮＴ）６２の代わりに、ガード区間制御部（ＧＩＮＳＣＮＴ）８１および電力制御部（ＰＣＮＴ）８２が設けられている。なお、ガード区間制御部８１および電力制御部８２の動作は、第４の実施例において説明した通りである。すなわち、ガード区間制御部８１は、対応する受信装置から送られてくる伝送距離情報Ｌに基づいて、挿入すべきガード区間の長さを決定する。また、電力制御部８２は、対応する受信装置から送られてくる伝送宇距離情報Ｌに基づいて、利得係数αを決定する。

図３９は、第８の実施例の受信装置の構成図である。この受信装置は、図３６に示した第７の実施例の受信装置の遅延差検出部１１１、ガード区間制御部１１２の代わりに、距離推定部（ＬＭＥＳ）１２１、変換テーブル（ＴＢＬ）１２２、ガード区間制御部（ＧＣＮＴ）１２３を備える。ここで、距離推定部１２１およびガード区間制御部１２３は、まず、第７の実施例と同様に、最適なガード区間長を決定する。その後、距離推定部１２１は、変換テーブル１２２にアクセスし、決定したガード区間長に対応する伝送距離を取得する。そして、その伝送距離を表す伝送距離情報Ｌを送信装置に通知する。なお、変換テーブル１２２は、図２８に示した変換テーブル８５に相当し、ガード区間長と伝送距離との対応関係が格納されている。

図４０は、図３９に示す距離推定部１２１の動作を示すフローチャートである。図４０において、ステップＳ１〜Ｓ７は、図３７に示した第７の実施例における処理と同じである。すなわち、ステップＳ１〜Ｓ７において、受信装置に設定すべきガード区間長τiが決定される。続いて、ステップＳ１１では、変換テーブル１１２を参照して、ガード区間長τi を伝送情報Li に変換する。そして、ステップＳ１２において、ステップＳ１１で取得した伝送情報を送信装置に通知する。

このように、本発明によれば、セルラ通信システムにおける基地局とそのサービスエリア内の移動機との間の伝送路で生じる最大伝送遅延差に応じてガード区間および送信電力が適切に設定されるので、干渉の発生が低減される。あるいは、伝送路の送信帯域内での伝送容量が最適化されるので、通信システムの効率的な運用が可能となり、総伝送容量を増加させることができる。

なお、ガード区間および送信電力は、送信装置と受信装置との間の回線の最大伝送遅延差（または、伝送距離）に応じて動的に制御されてもよいし、固定的に設定されてもよい。例えば、通信の開始時にガード区間および送信電力が決定され、以降、その通信が終了するまでそれらが変化しないようにしてもよい。また、通信中に、随時、ガード区間および送信電力が動的に調整されてもよい。さらに、送信装置および受信装置の位置が変化しない場合には、初期設定処理においてガード区間および送信電力が決定されてもよい。

また、本発明では、最大伝送遅延差（または、伝送距離）に応じてガード区間および送信電力が決定されるが、ガード区間長と送信電力の関係は、たとえば、実験またはシミュレーション等により予め一意に決められていてもよい。

請求の範囲

1.　直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する通信システムであって、

上記送信装置は、

信号系列を用いて複数の副搬送波を変調する変調手段と、

上記変調手段の出力にガード区間を挿入する挿入手段と、

上記ガード区間が挿入された変調信号を送信する送信手段を有し、

上記受信手段は、

上記送信装置から送信された変調信号について副搬送波ごとにガード区間の削除処理と復調処理を行い、信号系列を再生する復調手段を有し、

上記ガード区間の長さは、上記送信装置と上記受信装置との間の通信環境に基づいて決定される通信システム。

2.　請求項1に記載の通信システムであって、

上記送信装置は、上記ガード区間の長さに応じて上記変調信号を送信する際の送信電力を制御する電力制御手段をさらに有する。

3.　直交周波数分割多重を利用して送信装置から第1の受信装置を含む複数の受信装置へ信号を伝送する通信システムであって、

上記送信装置は、

第1の受信装置へ伝送する第1の信号系列、および第1の受信装置とは異なる他の受信装置へ伝送する第2の信号系列が多重された信号系列を用いて複数の副搬送波を変調する変調手段と、

上記第1の信号系列の変調出力に第1のガード区間を、上記第2の信号系列の変調出力に第2のガード区間をそれぞれ挿入する挿入手段と、

上記第1のガード区間と第2のガード区間がそれぞれ挿入された変調信号

を送信する送信手段を有し、

上記第１の受信装置は、

上記第１のガード区間の削除処理と復調処理を行い、第１の信号系列を再生する復調手段を有し、

上記第１のガード区間の長さは、上記送信装置と上記第１の受信装置との間の通信環境に基づいて決定されるとともに、上記第２のガード区間の長さは、上記送信装置と上記他の受信装置との間の通信環境に基づいて決定される。

４．　直交周波数分割多重を利用して送信装置から第１の受信装置を含む複数の受信装置へ信号を伝送する通信システムであって、

上記送信装置は、

第１の受信装置へ伝送する第１の信号系列、および上記送信装置の通信エリア内の第１の受信装置を含む複数の受信装置に伝送する第２の信号系列が多重された信号系列を用いて複数の副搬送波を変調する変調手段と、

上記第１の信号系列の変調出力に第１のガード区間を、上記第２の信号系列の変調出力に第２のガード区間をそれぞれ挿入する挿入手段と、

上記第１のガード区間と第２のガード区間がそれぞれ挿入された変調信号を送信する送信手段を有し、

上記第１の受信装置は、

上記第１のガード区間の削除処理と第２のガード区間の削除処理と復調処理を行い、第１の信号系列と第２の信号系列を再生する復調手段を有し、

上記第１のガード区間の長さは、上記送信装置と上記第１の受信装置との間の通信環境に基づいて決定される

５．　請求項４に記載の通信システムであって、

上記第２ガード区間の長さは、通信エリア内に存在する複数の受信装置が第２の信号系列を再生できるように決定される。

６．　請求項１に記載の通信システムであって、

上記受信装置は、上記送信装置と当該受信装置との間の回線の最大伝送遅延差を検出する検出手段をさらに有し、

上記挿入手段は、上記検出手段により検出された最大伝送遅延差に基づいて決まる長さのガード区間を挿入し、

上記削除手段は、その最大伝送遅延差に従ってガード区間を削除する。

７．　請求項１に記載の通信システムであって、

上記受信装置は、上記送信装置と当該受信装置との間の伝送距離を推定する推定手段をさらに有し、

上記挿入手段は、上記推定手段により推定された伝送距離に基づいて決まる長さのガード区間を挿入し、

上記削除手段は、その推定された伝送距離に従ってガード区間を削除する。

８．　請求項１に記載の通信システムであって、

上記送信装置は、上記送信装置と当該受信装置との間の伝送距離を推定する推定手段をさらに有し、

上記挿入手段は、上記推定手段により推定された伝送距離に基づいて決まる長さのガード区間を挿入し、

上記削除手段は、その推定された伝送距離に従ってガード区間を削除する。

９．　請求項８に記載の通信システムであって、

上記推定手段は、当該送信装置から信号が送信されたときから、上記受信装置からその信号に対応する応答が返ってくるまでの時間に基づいて上記伝送距離を推定する。

１０．請求項１に記載の通信システムであって、

上記受信装置は、上記送信装置から当該受信装置へ信号が伝送されたときの通信品質をモニタするモニタ手段をさらに有し、

上記ガード区間の長さは、予め決められた所定の通信品質が満たされるように決定される。

１１． 直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する方法であって、

直交周波数分割多重を利用して信号系列を変調し、

上記変調により得られた信号に対して、上記送信装置と上記受信装置との間の通信環境に基づいて決定される長さのガード区間を挿入し、

上記ガード区間が挿入された変調信号を送信する信号伝送方法。

１２． 請求項１１に記載の方法であって、

上記ガード区間が挿入された変調信号は、そのガード区間の長さに応じてその送信電力が制御される。

１３． 直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する方法であって、

送信相手先の異なる第１の信号系列および第２の信号系列を直交周波数多重を利用して変調し、

上記変調された第１の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第１のガード区間を挿入し、上記変調された第２の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第２のガード区間を挿入し、

上記第１のガード区間および第２のガード区間がそれぞれ挿入された変調信号を送信する信号伝送方法。

１４． 請求項１１に記載の方法であって、

上記ガード区間の長さは、上記送信装置と上記受信装置との間の回線の最大伝送遅延差または伝送距離に基づいて決定される。

１５． 請求項１１に記載の方法であって、

上記送信装置から上記受信装置へ信号が伝送されたときの通信品質をモニタし、

上記ガード区間の長さは、予め決められた所定の通信品質が満たされるように決定される。

１６．セルラ通信システムにおいて直交周波数分割多重を利用して移動機へ信号を伝送する基地局装置であって、

直交周波数分割多重を利用して信号系列を変調する変調手段と、

上記変調手段により得られた変調信号に対して、当該基地局装置と上記信号系列を送信すべき移動機との間の通信環境に基づいて決定される長さのガード区間を挿入する挿入手段と、

上記ガード区間が挿入された変調信号を送信する送信手段と、

を有する基地局装置。

１７．請求項１６に記載の基地局装置であって、

上記ガード区間の長さに応じて上記変調信号を送信する際の送信電力を制御する電力制御手段をさらに有する。

１８．セルラ通信システムにおいて直交周波数分割多重を利用して移動機へ信号を伝送する基地局装置であって、

送信相手先の異なる第１の信号系列および第２の信号系列を直交周波数多重を利用してそれぞれ変調する変調手段と、

上記変調手段により得られた変調された第１の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第１のガード区間を挿入するとともに、上記変調手段により得られた変調された第２の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第２のガード区間を挿入する挿入手段と、

上記第１のガード区間および第２のガード区間がそれぞれ挿入された変調信

号を送信する送信手段と、

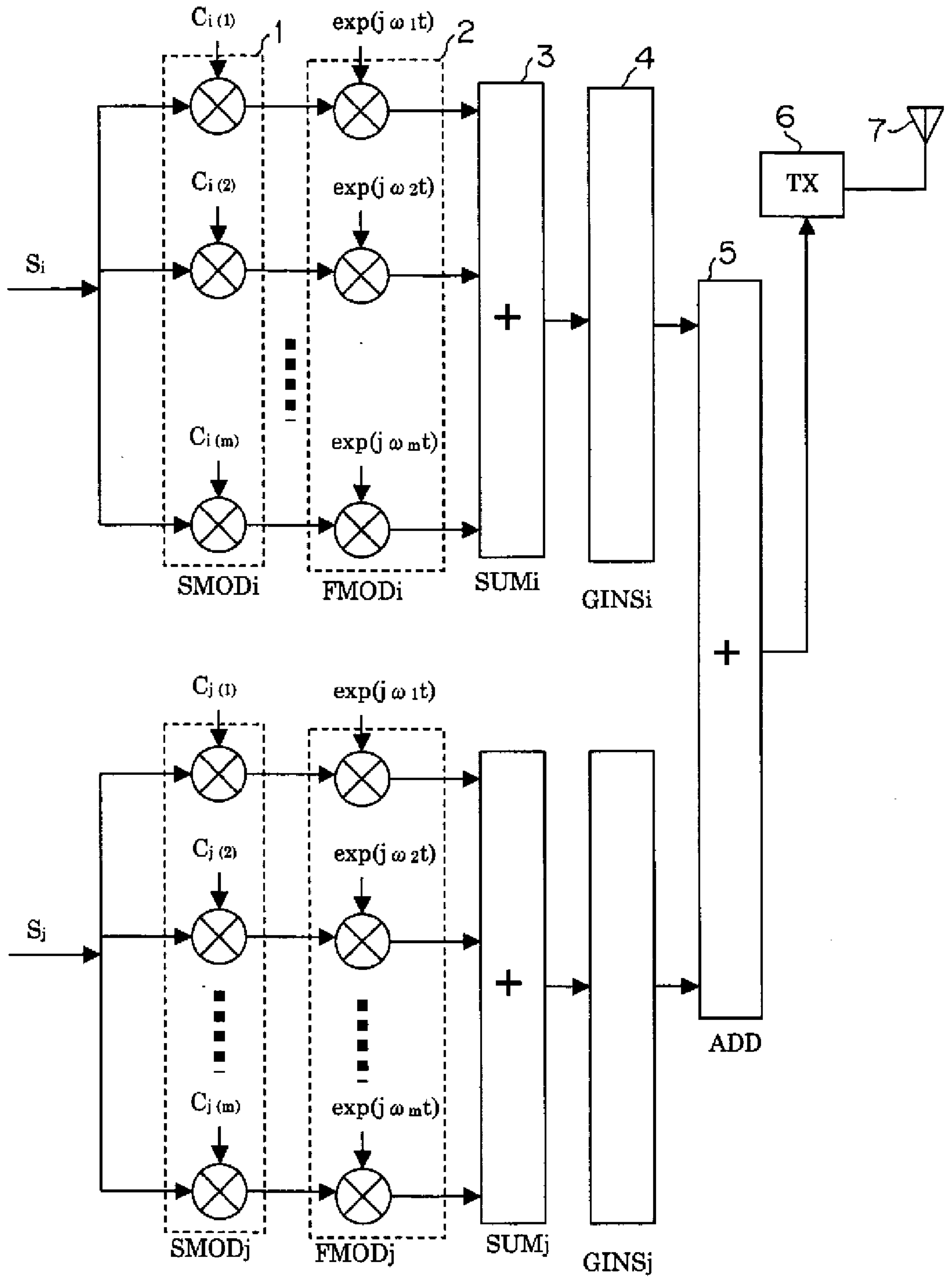
を有する基地局装置。

１９． セルラ通信システムにおいて、直交周波数分割多重を利用して基地局から送信された信号を受信する移動機であって、

受信した信号が自移動機宛ての第１の信号系列と他移動機宛ての第２の信号系列を含む場合、第１の信号系列に対応する第１のガード区間の削除処理と復調処理を行う復調手段を有する移動機。

２０． セルラ通信システムにおいて、直交周波数分割多重を利用して基地局から送信された信号を受信する移動機であって、

受信した信号が自移動機宛ての第１の信号系列と自移動機を含む複数の移動機宛ての第２の信号系列を含む場合、第１の信号系列に対応する第１のガード区間の削除処理と、第２の信号系列に対応する第２のガード区間の削除処理と、復調処理を行う復調手段を有する移動機。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S_i
+
SMODi
FMODi
SUMi
GINSi
ADD
$C_{j(1)}$
$C_{j(2)}$
$C_{j(m)}$
S_j
SMODj
FMODj
SUMj
GINSj

図1

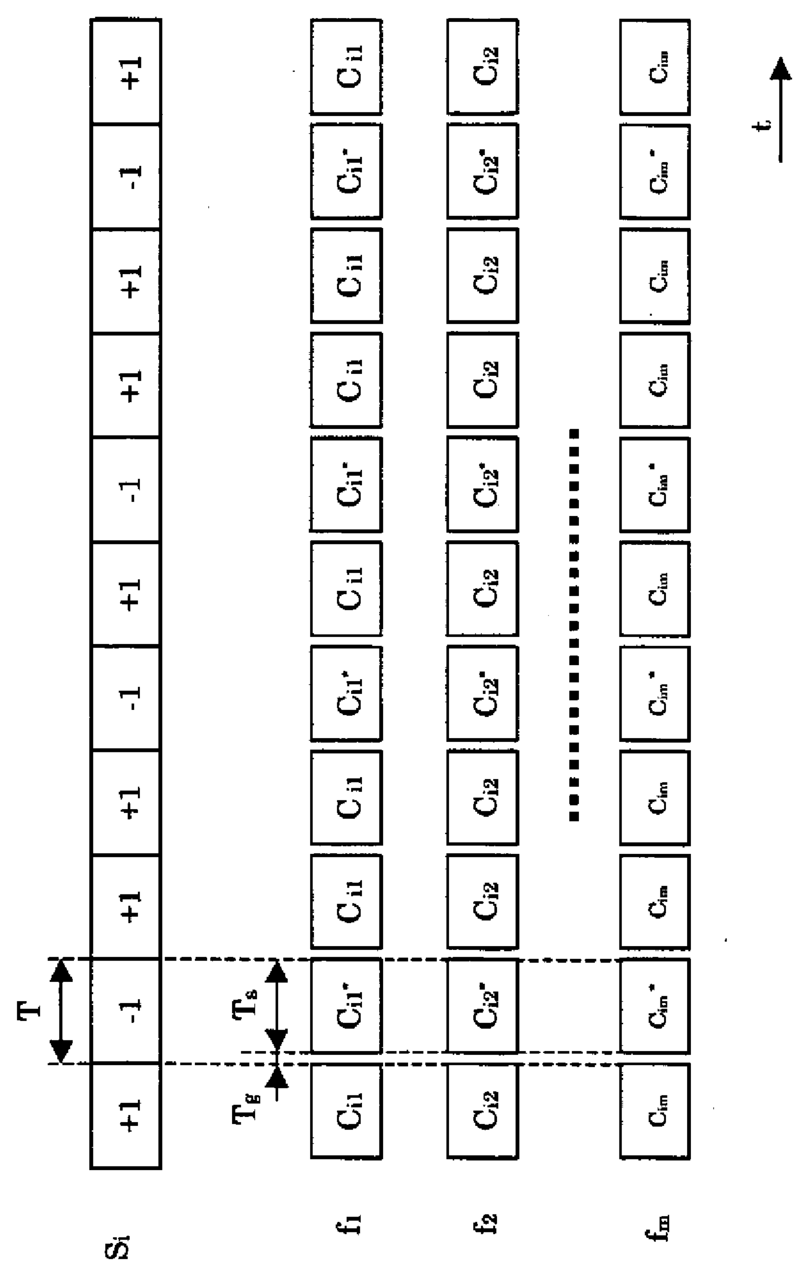


図2

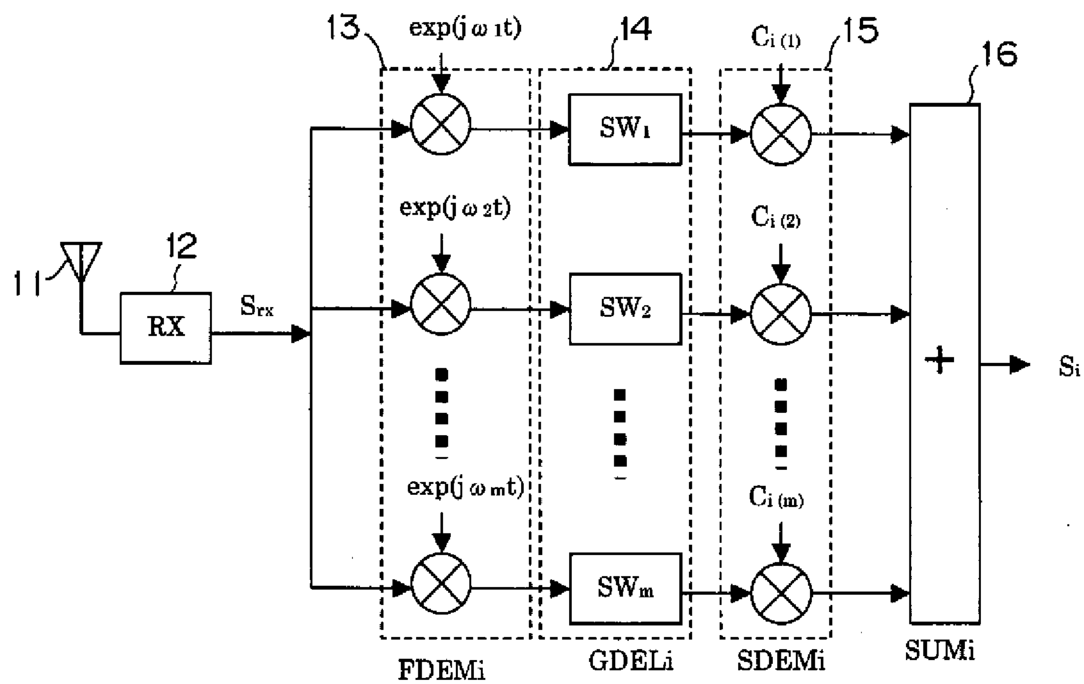

13
exp(jω1t)
14
Ci(1)
15
16
SW1
exp(jω2t)
Ci(2)
11
12
RX
Srx
SW2
+
Si
exp(jωmt)
Ci(m)
SWm
FDEMi
GDELi
SDEMi
SUMi

図3

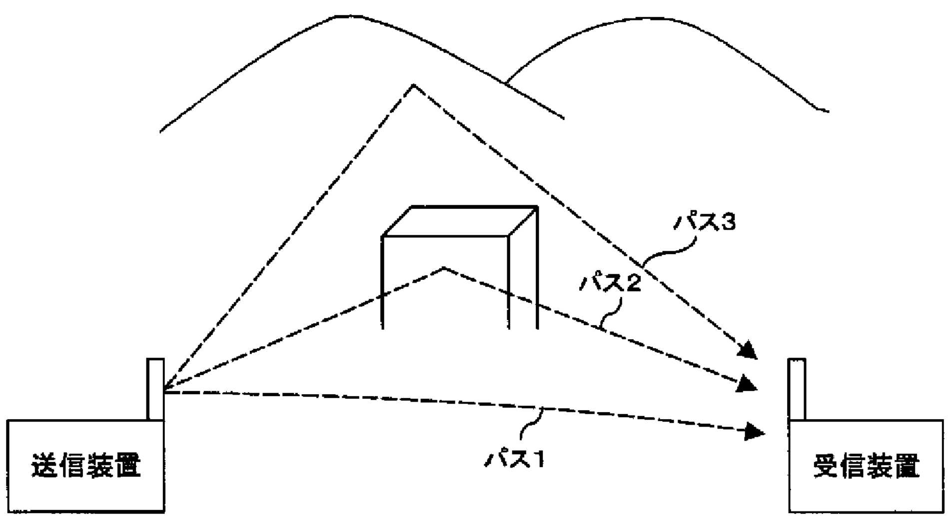


図4

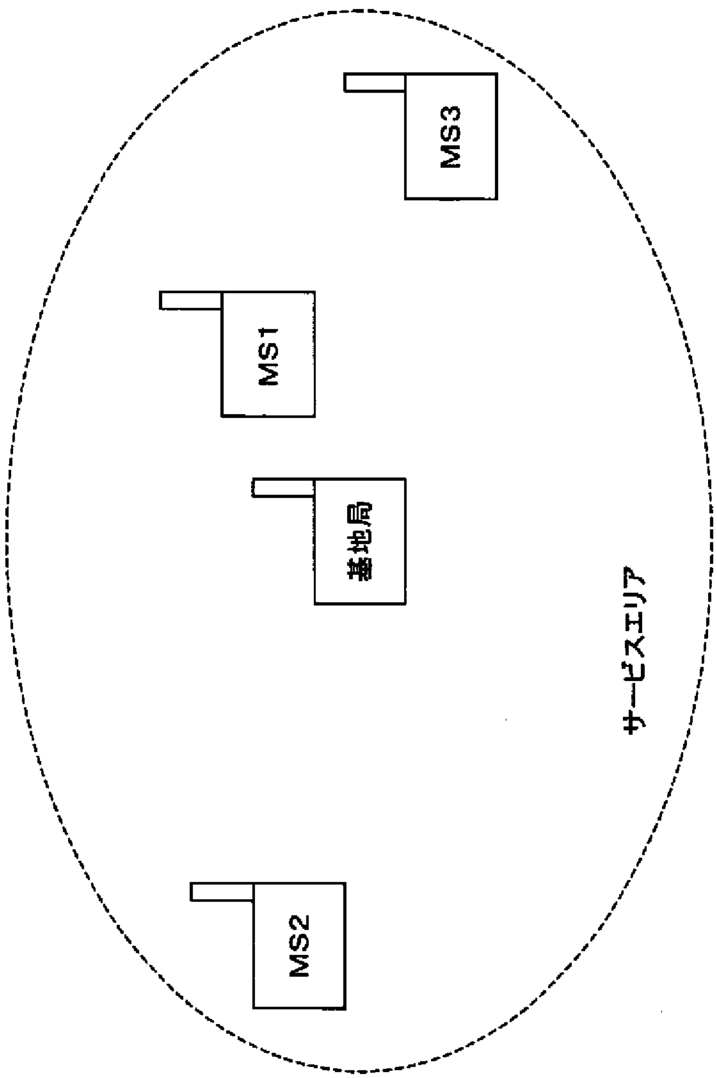


図5

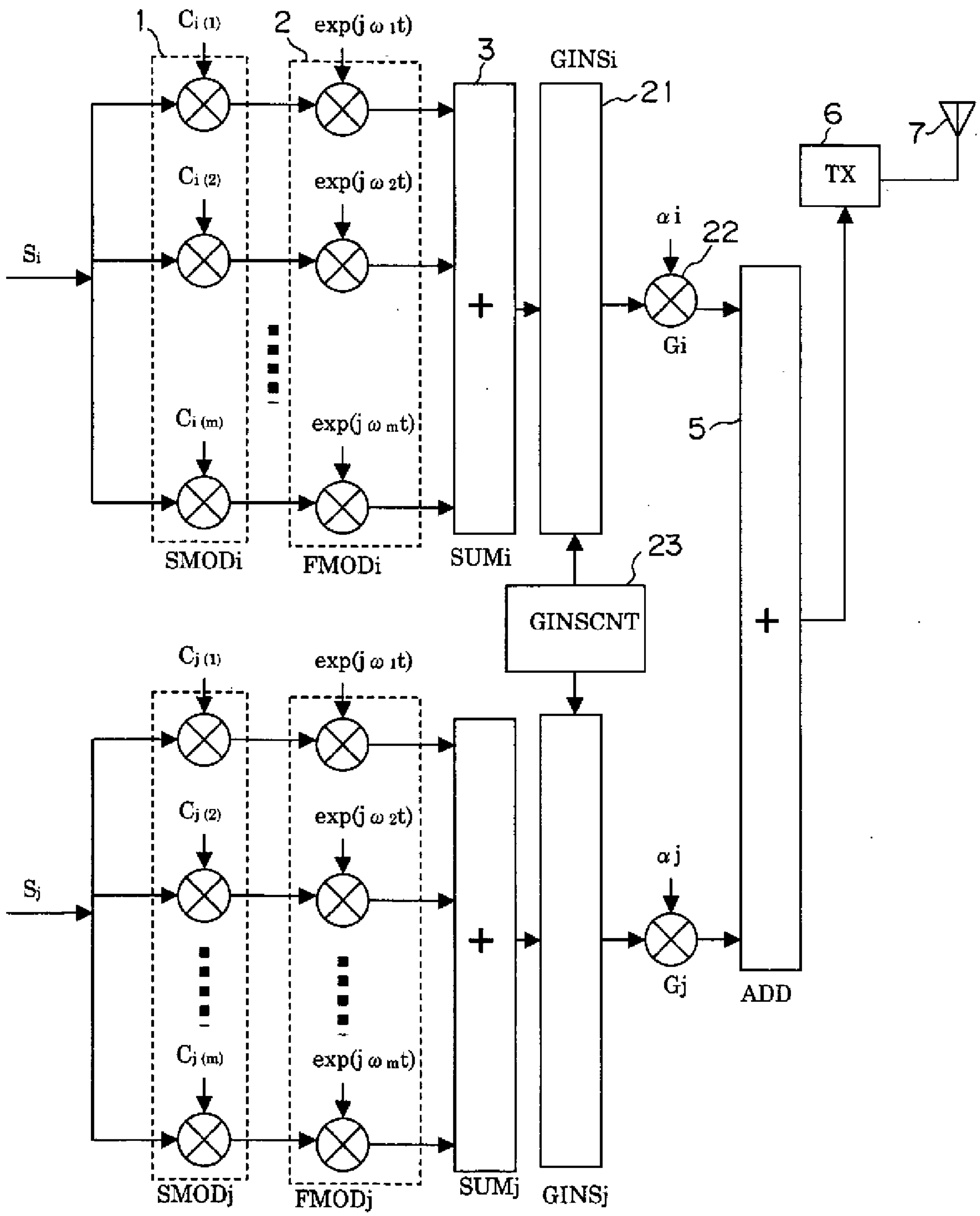

1
$C_{i(1)}$
2
exp(j ω 1t)
3
GINSi
21
6
7
TX
$C_{i(2)}$
exp(j ω 2t)
S_i
α i
22
+
Gi
5
$C_{i(m)}$
exp(j ω mt)
SMODi
FMODi
SUMi
23
GINSCNT
+
$C_{j(1)}$
exp(j ω 1t)
$C_{j(2)}$
exp(j ω 2t)
α j
S_j
+
Gj
ADD
$C_{j(m)}$
exp(j ω mt)
SMODj
FMODj
SUMj
GINSj

図6

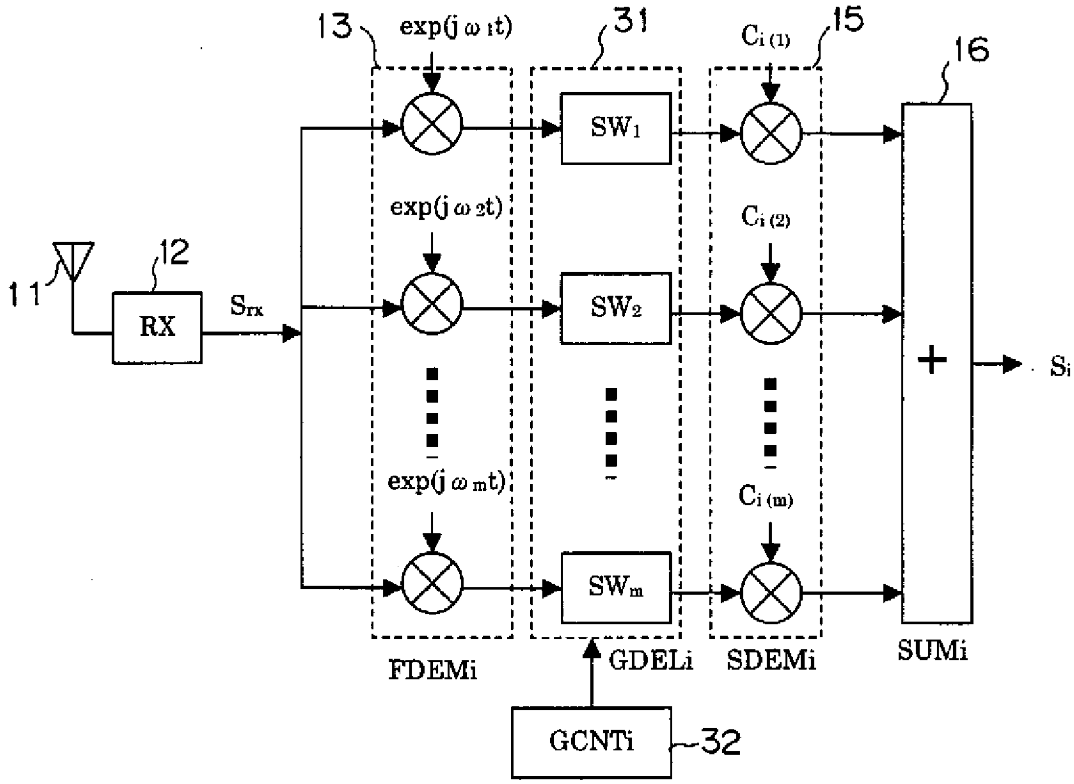

13
exp(jω1t)
31
Ci(1)
15
16
SW1
exp(jω2t)
Ci(2)
12
11
RX
Srx
SW2
+
Si
exp(jωmt)
Ci(m)
SWm
FDEMi
GDELi
SDEMi
SUMi
GCNTi
32

図7

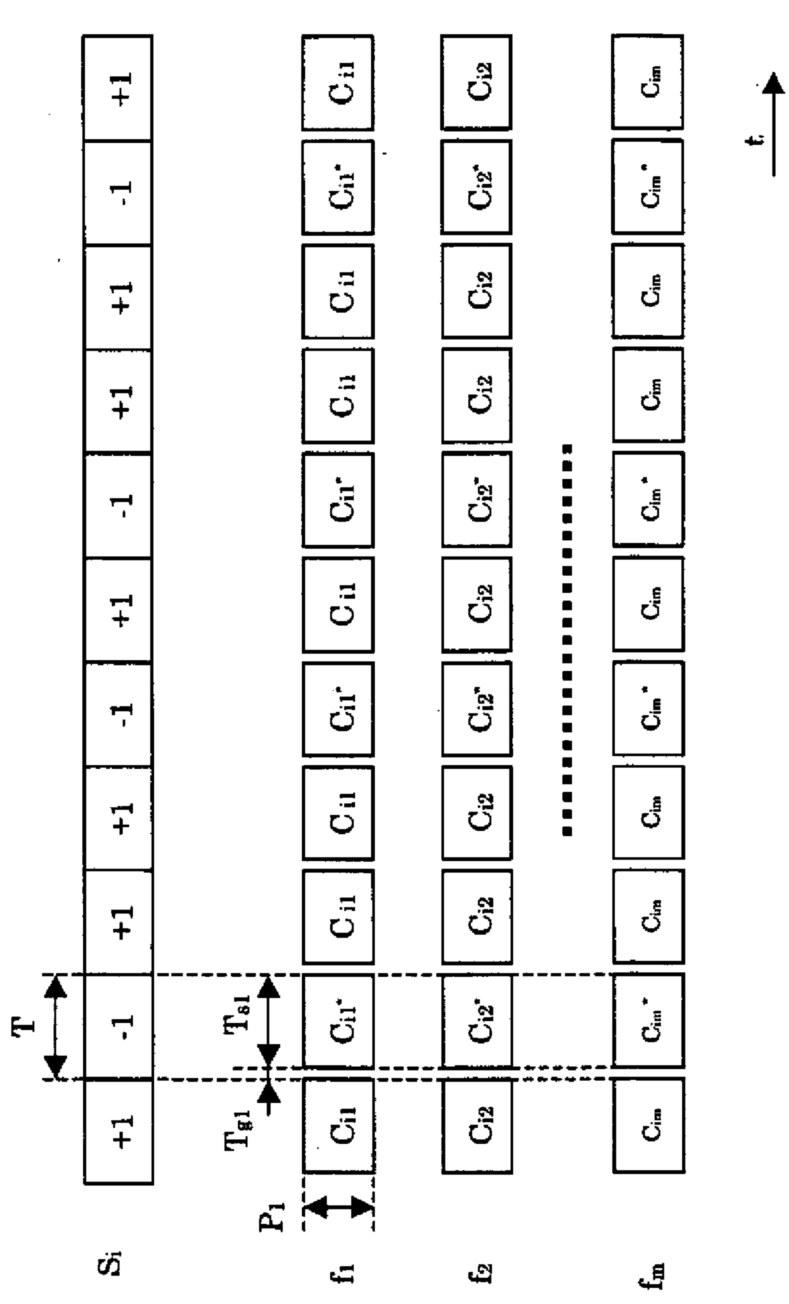


図8

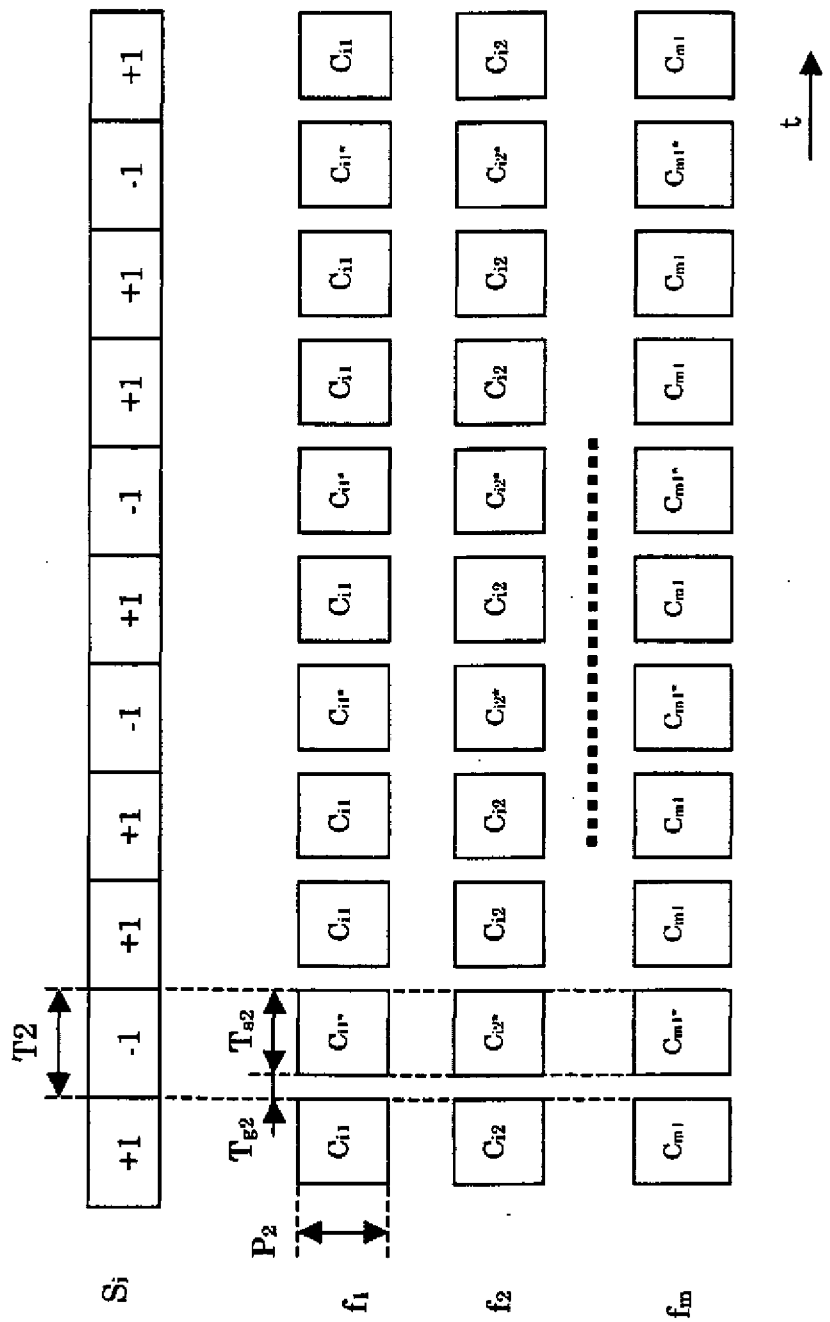


図9

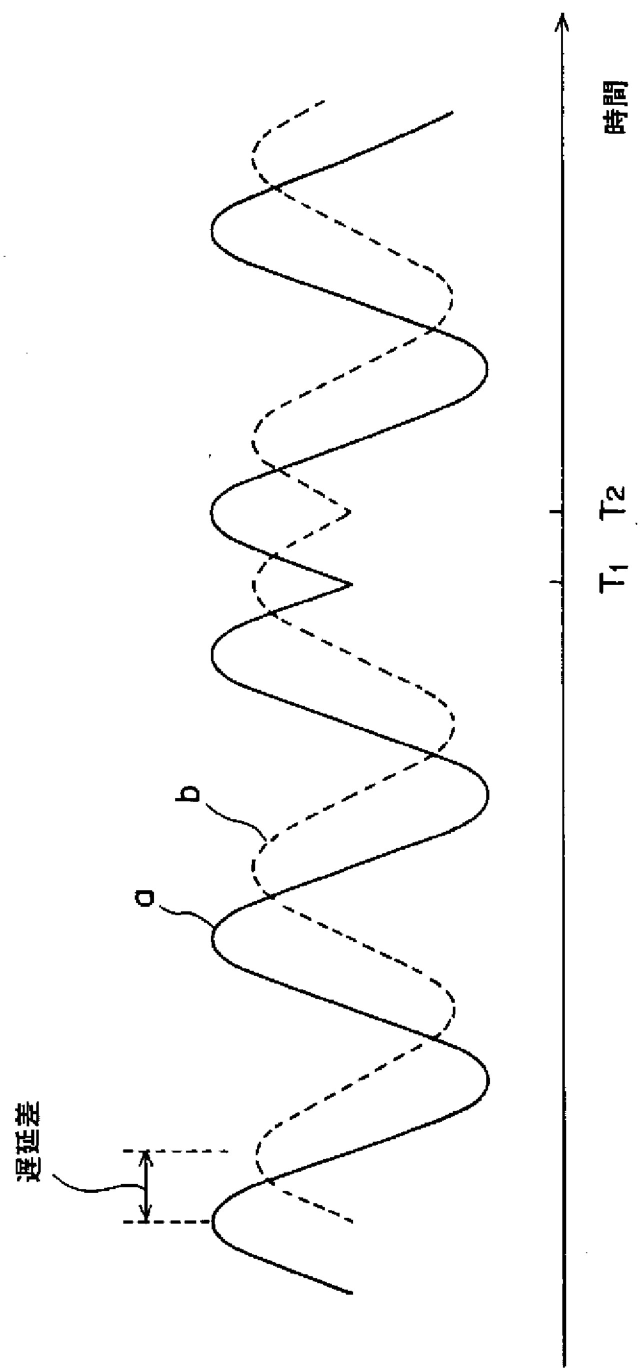


図10

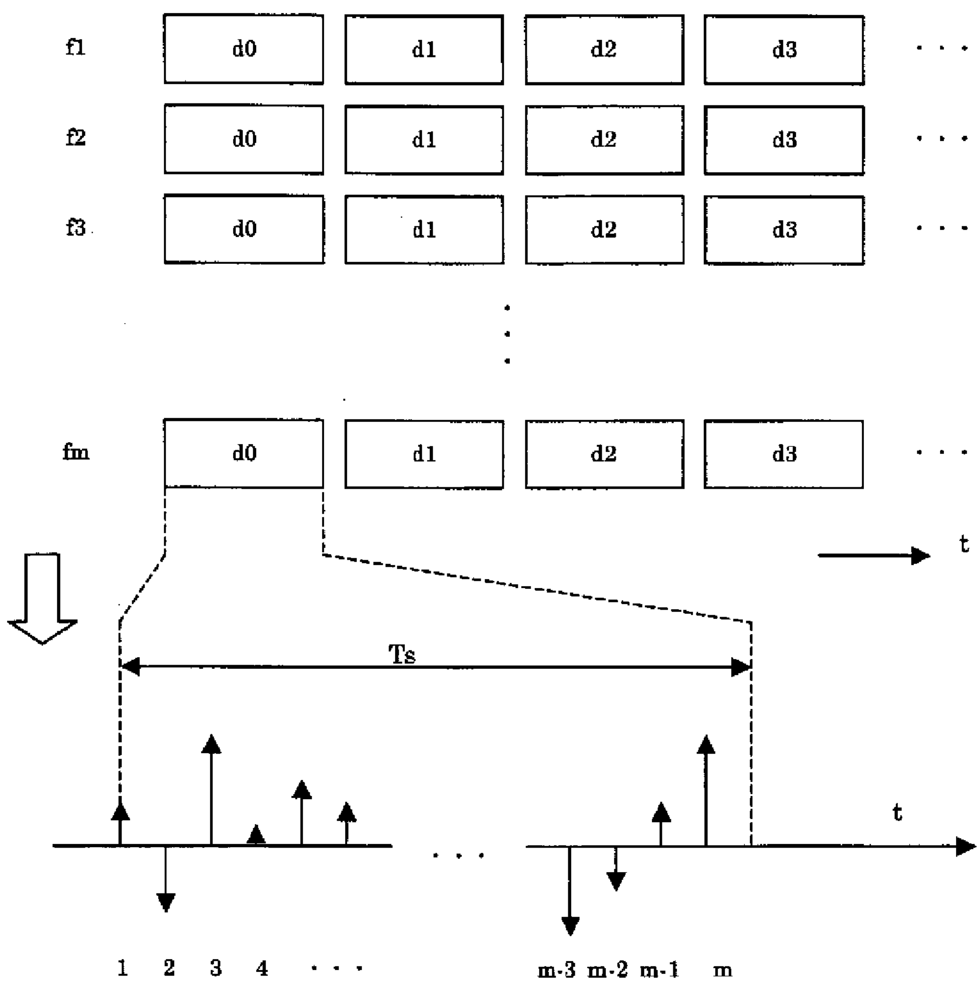


図11

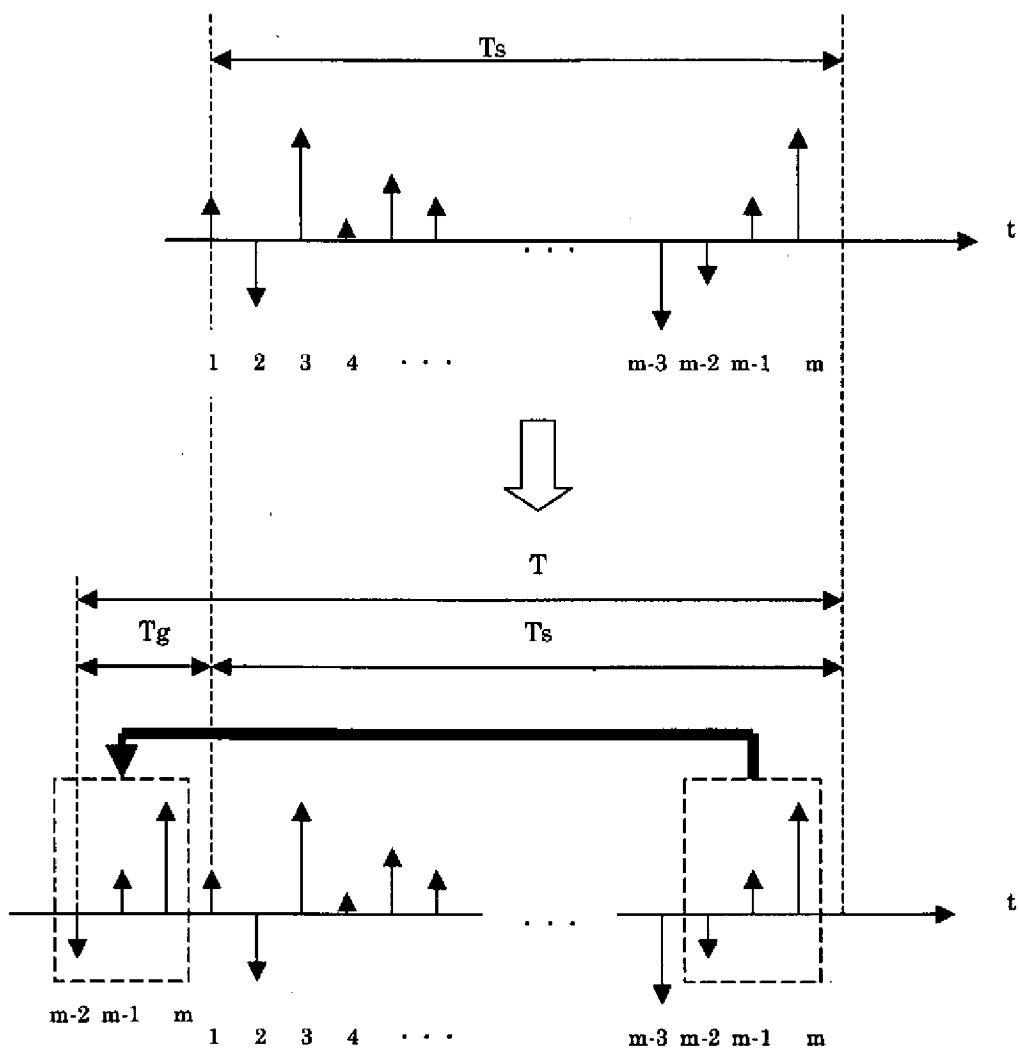


図12

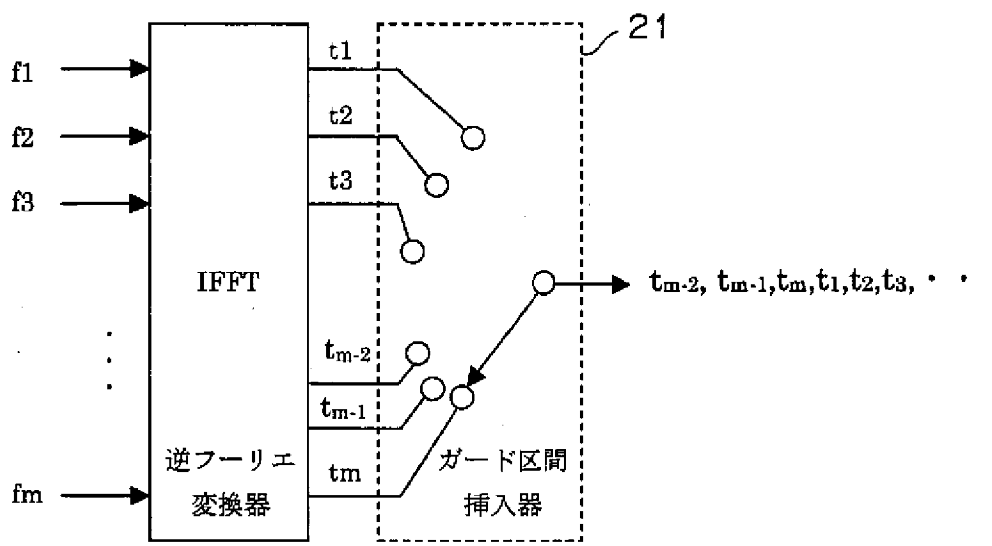

21
f1
f2
f3
fm
IFFT
逆フーリエ
変換器
t1
t2
t3
tm-2
tm-1
tm
ガード区間
挿入器
tm-2, tm-1,tm,t1,t2,t3, ・ ・ ・

図13

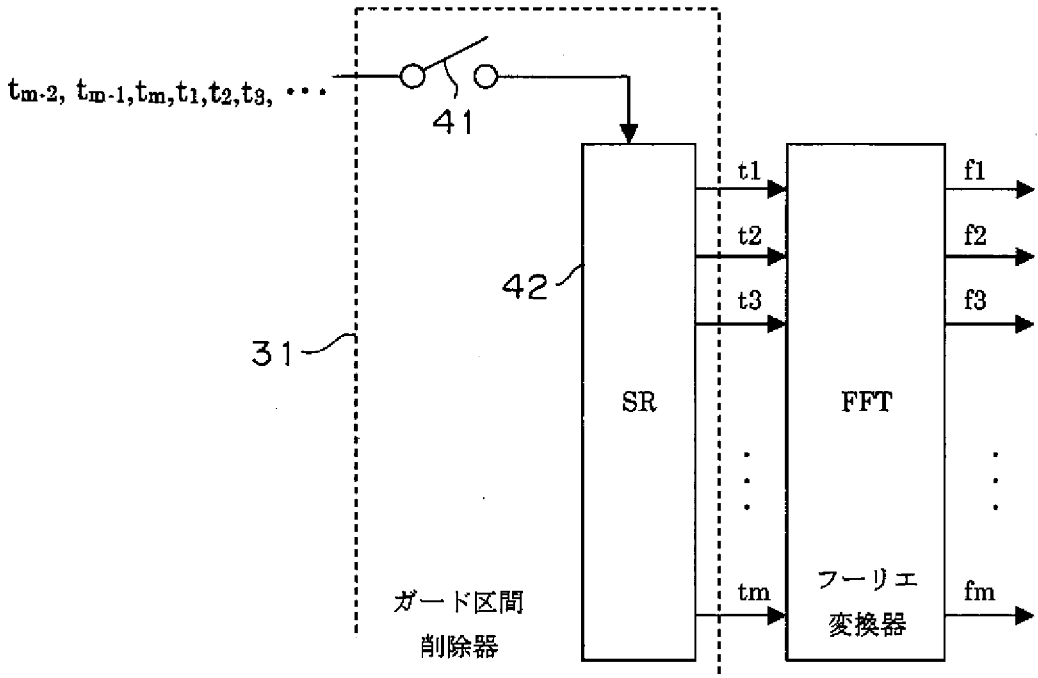

t_{m-2}, t_{m-1},t_m,t_1,t_2,t_3, ・・・
41
42
31
SR
FFT
t1
t2
t3
tm
f1
f2
f3
fm
ガード区間
削除器
フーリエ
変換器

図14

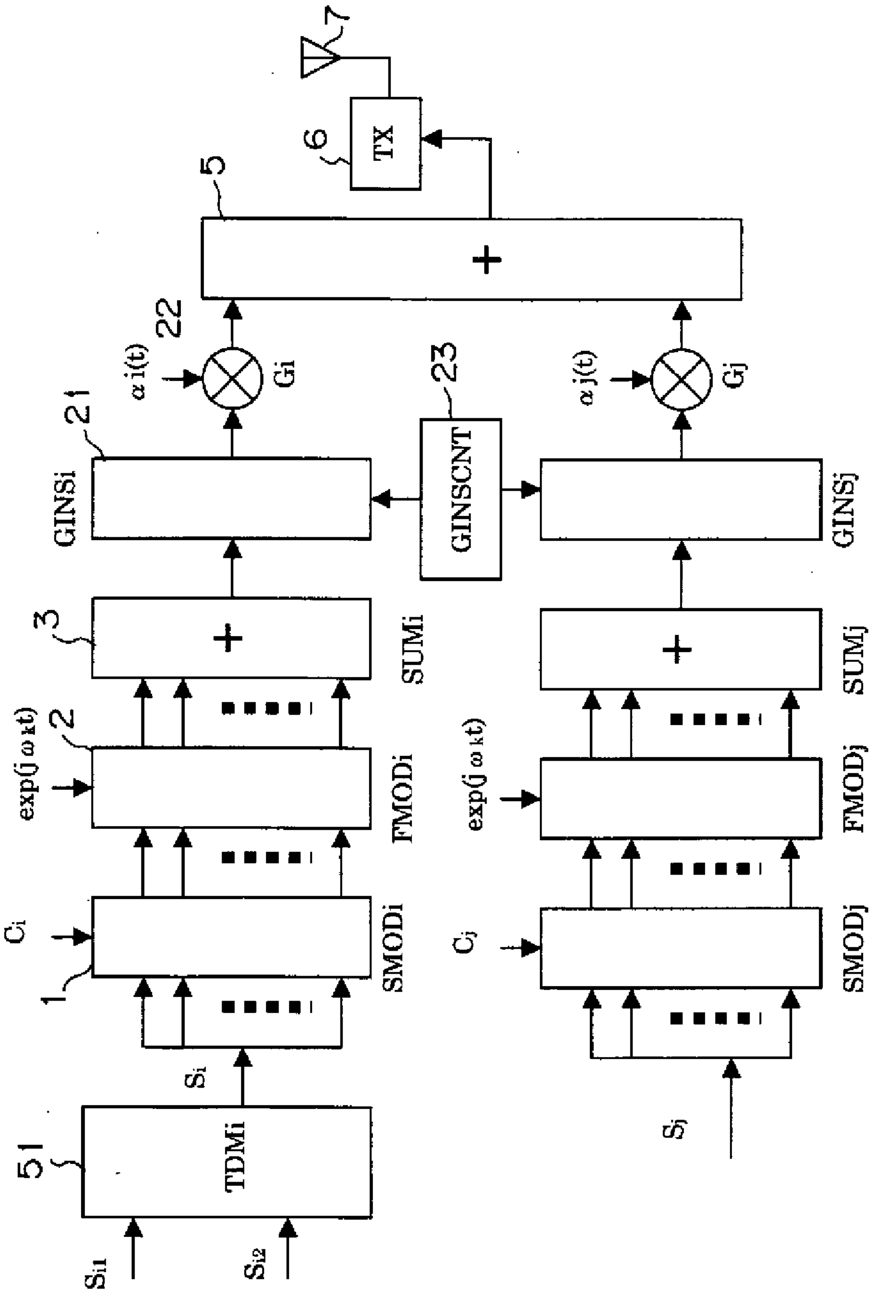

7
6
TX
5
+
22
α i(t)
Gi
23
GINSCNT
α j(t)
Gj
21
GINSi
GINSj
3
+
SUMi
SUMj
2
exp(jωkt)
FMODi
FMODj
C_i
C_j
1
SMODi
SMODj
S_i
S_j
51
TDMi
S_{i1}
S_{i2}

図15

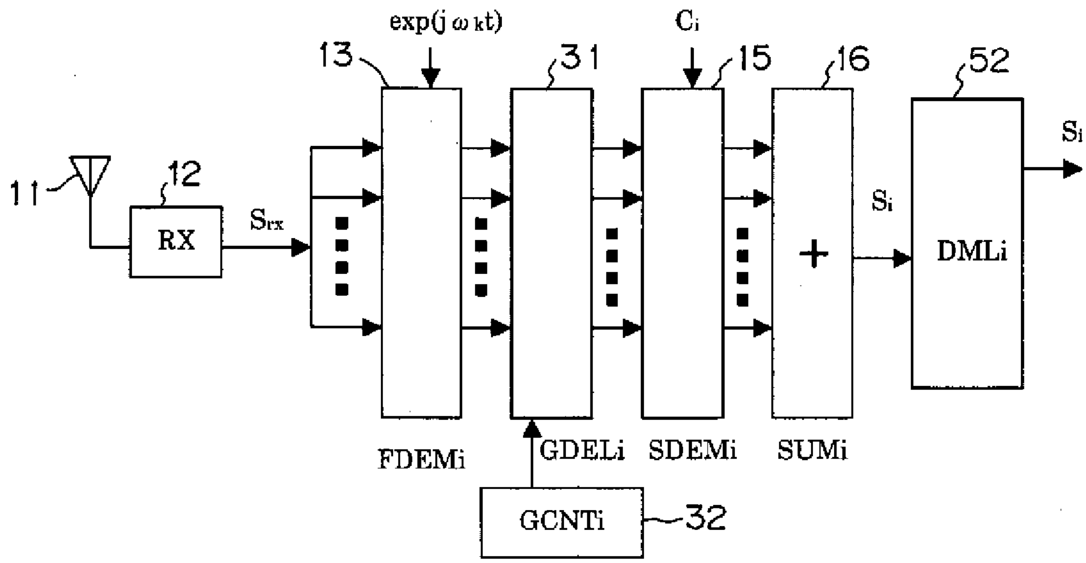
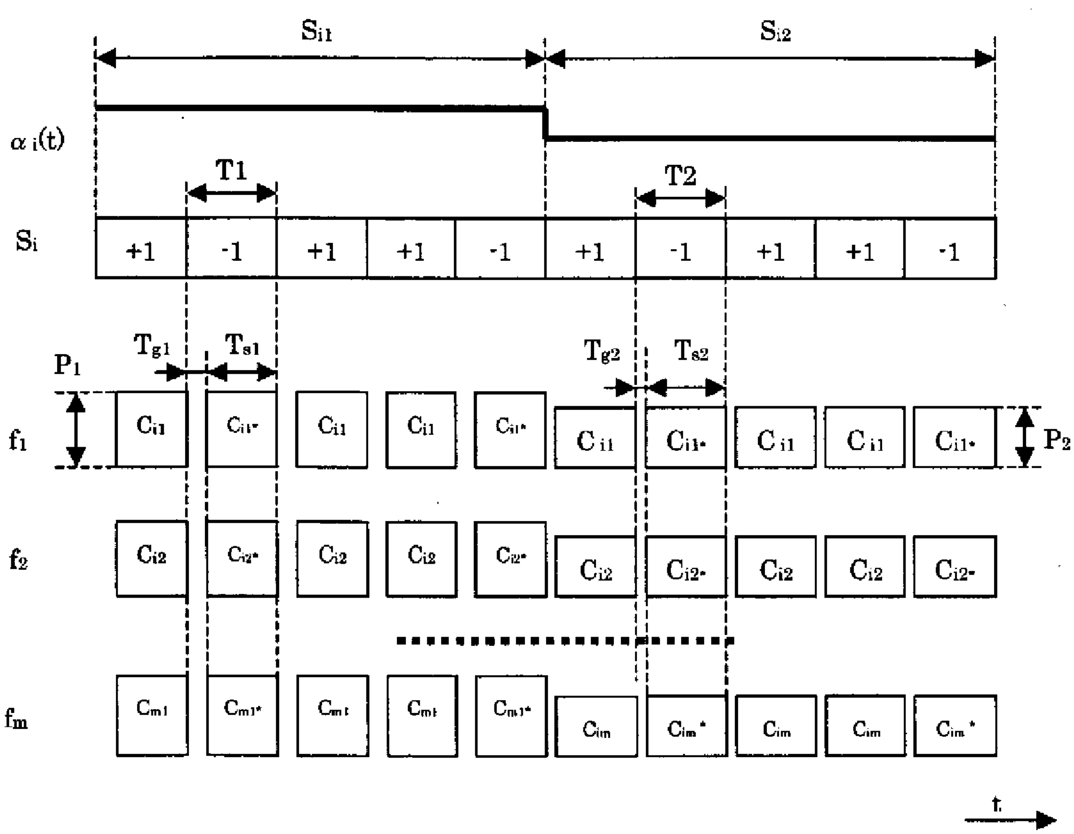
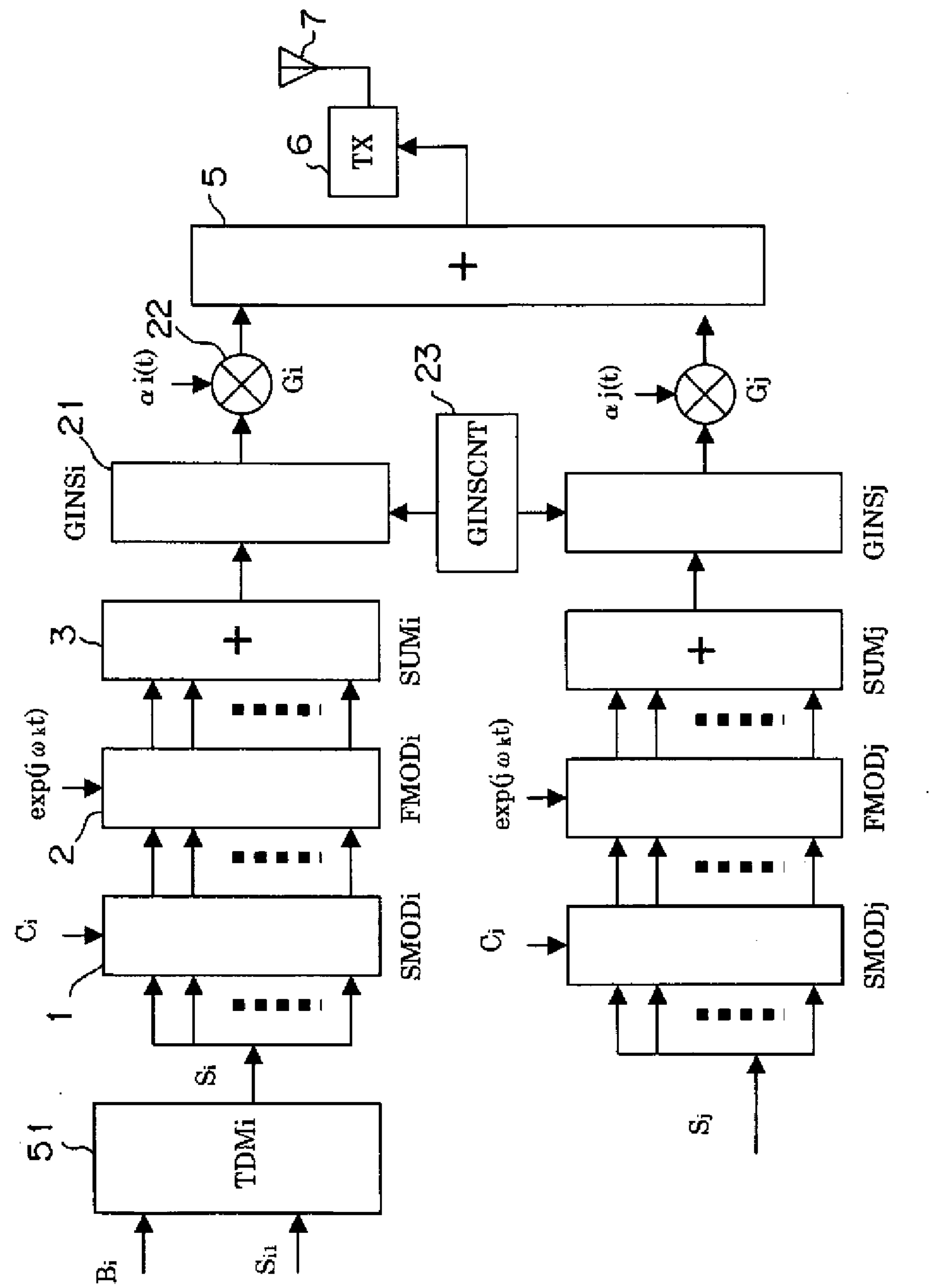

exp(jωkt)
13
31
Ci
15
16
52
Si1
11
12
RX
Srx
Si
+
DMLi
FDEMi
GDELi
SDEMi
SUMi
GCNTi
32

図16



S_{i1}
S_{i2}
$\alpha_i(t)$
T1
T2
S_i
+1
-1
+1
+1
-1
+1
-1
+1
+1
-1
P_1
T_{g1}
T_{s1}
T_{g2}
T_{s2}
P_2
f_1
f_2
f_m
C_{i1}
C_{i1^*}
C_{i2}
C_{i2^*}
C_{m1}
C_{m1^*}
C_{im}
C_{im^*}
t

図17



TDMi
TX
GINSCNT
GINSi
GINSj
SUMi
SUMj
FMODi
FMODj
SMODi
SMODj
Gi
Gj

図18

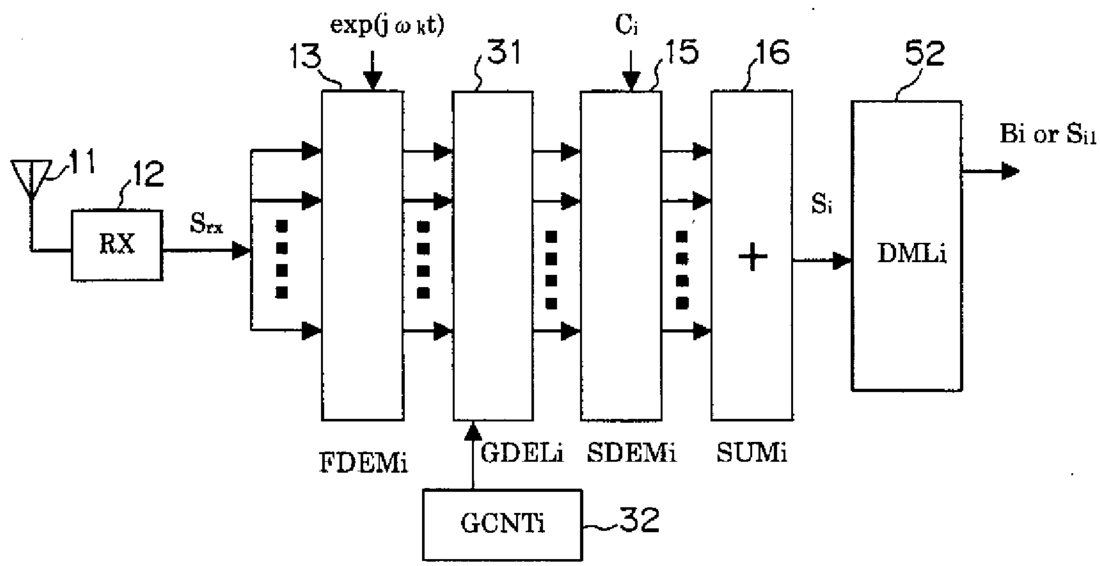

exp(j ω kt)
13
31
Ci
15
16
52
11
12
RX
Srx
Bi or Si1
Si
+
DMLi
FDEMi
GDELi
SDEMi
SUMi
GCNTi
32

図19

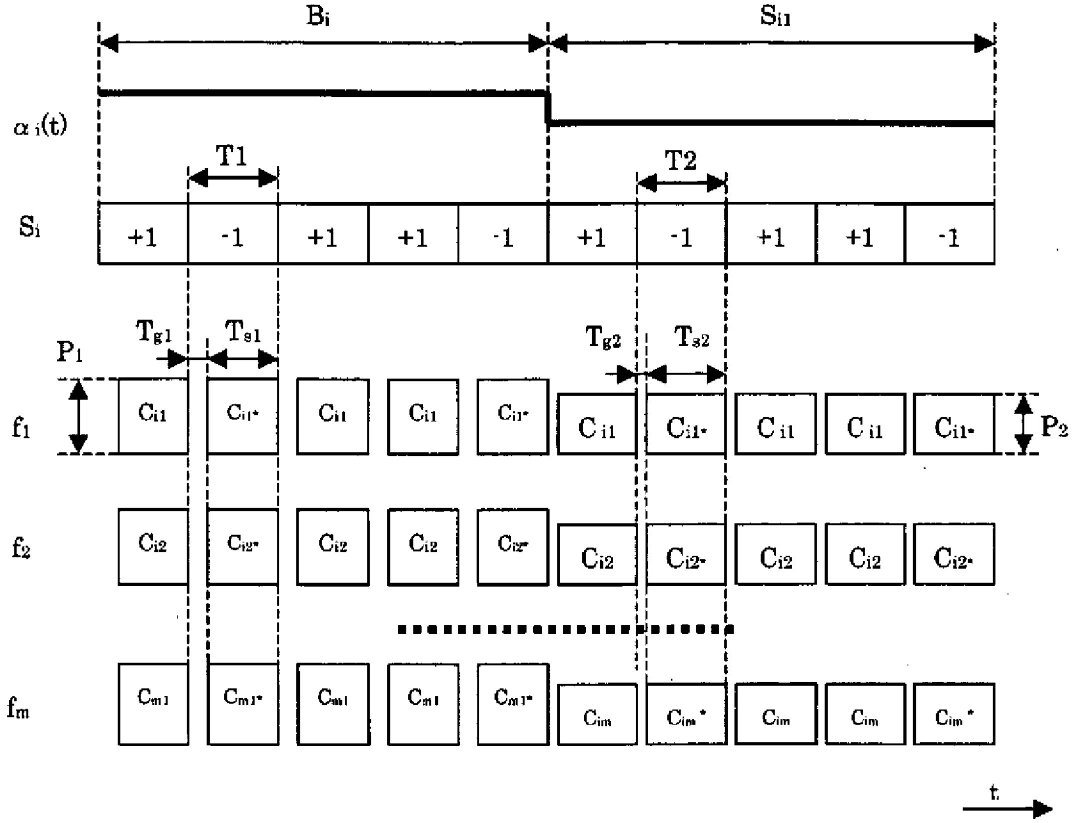


図20

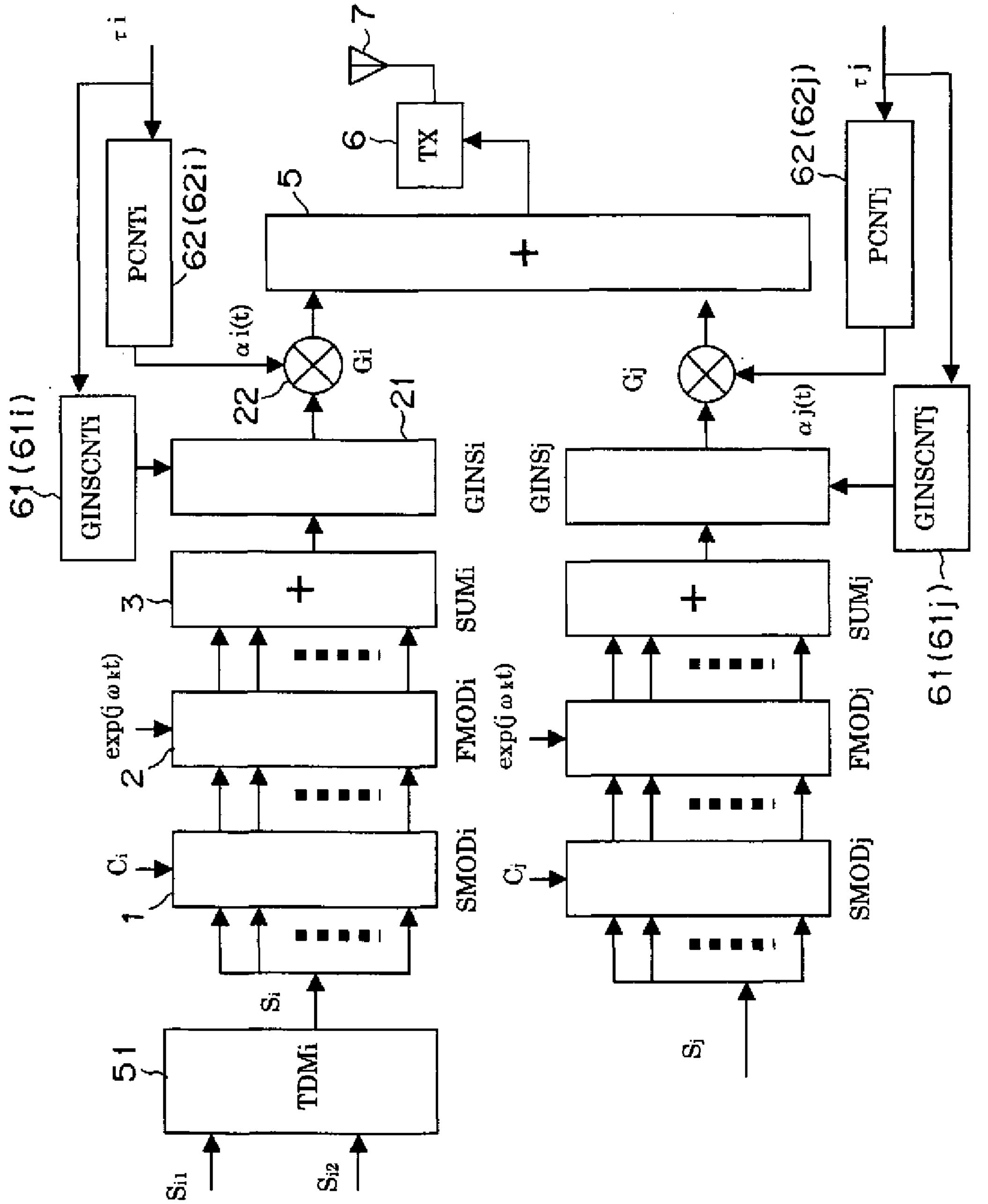

τi
62(62i)
PCNTi
61(61i)
GINSCNTi
αi(t)
22
Gi
21
GINSi
3
SUMi
exp(jωkt)
2
FMODi
Ci
1
SMODi
Si
51
TDMi
Si1
Si2
7
6
TX
5
τj
62(62j)
PCNTj
Gj
αj(t)
GINSCNTj
61(61j)
GINSj
SUMj
FMODj
Cj
SMODj
Sj

図21

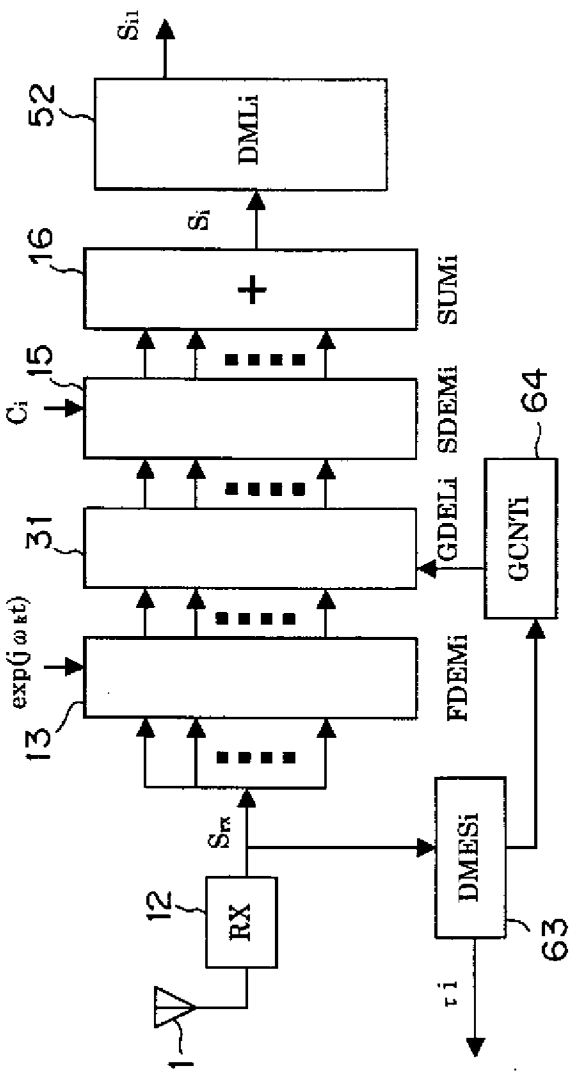

DMLi
SUMi
SDEMi
GDELi
FDEMi
GCNTi
DMESi
RX
S_{i1}
S_i
C_i
$\exp(j\omega_k t)$
S_{rx}
τi
11
12
13
15
16
31
52
63
64

図22

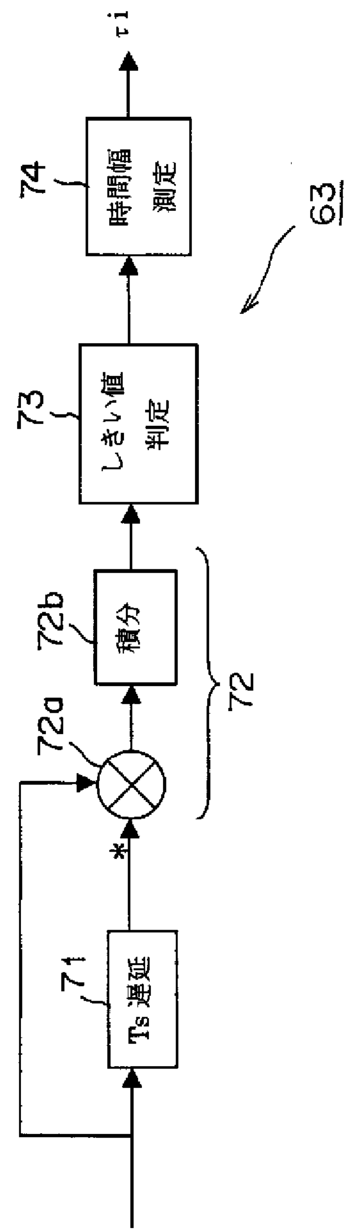


図23

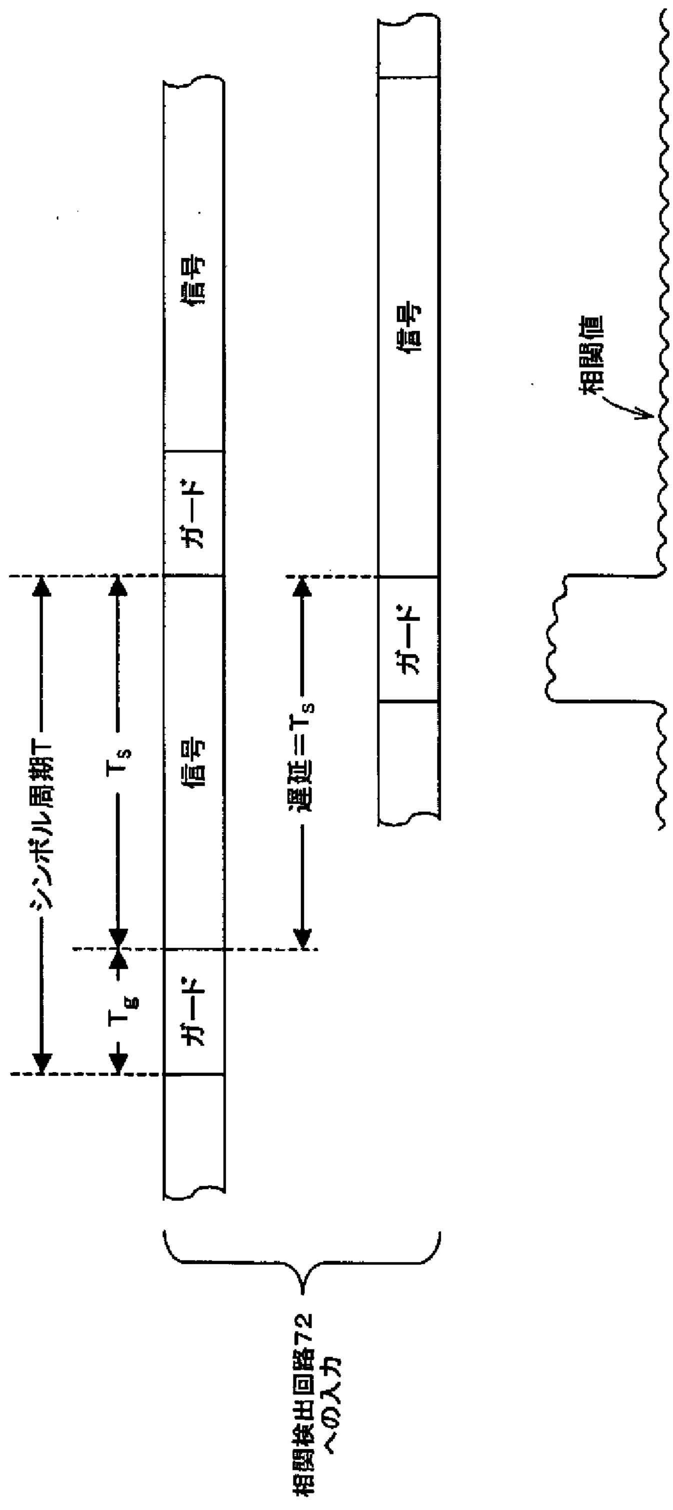


図24

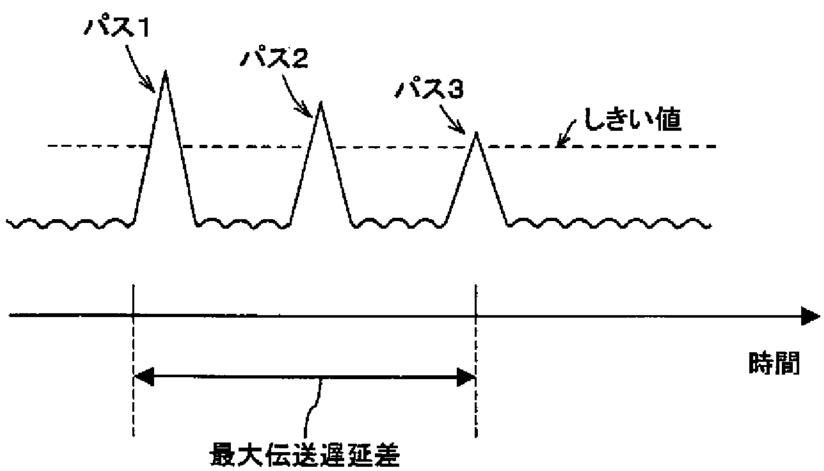


図25

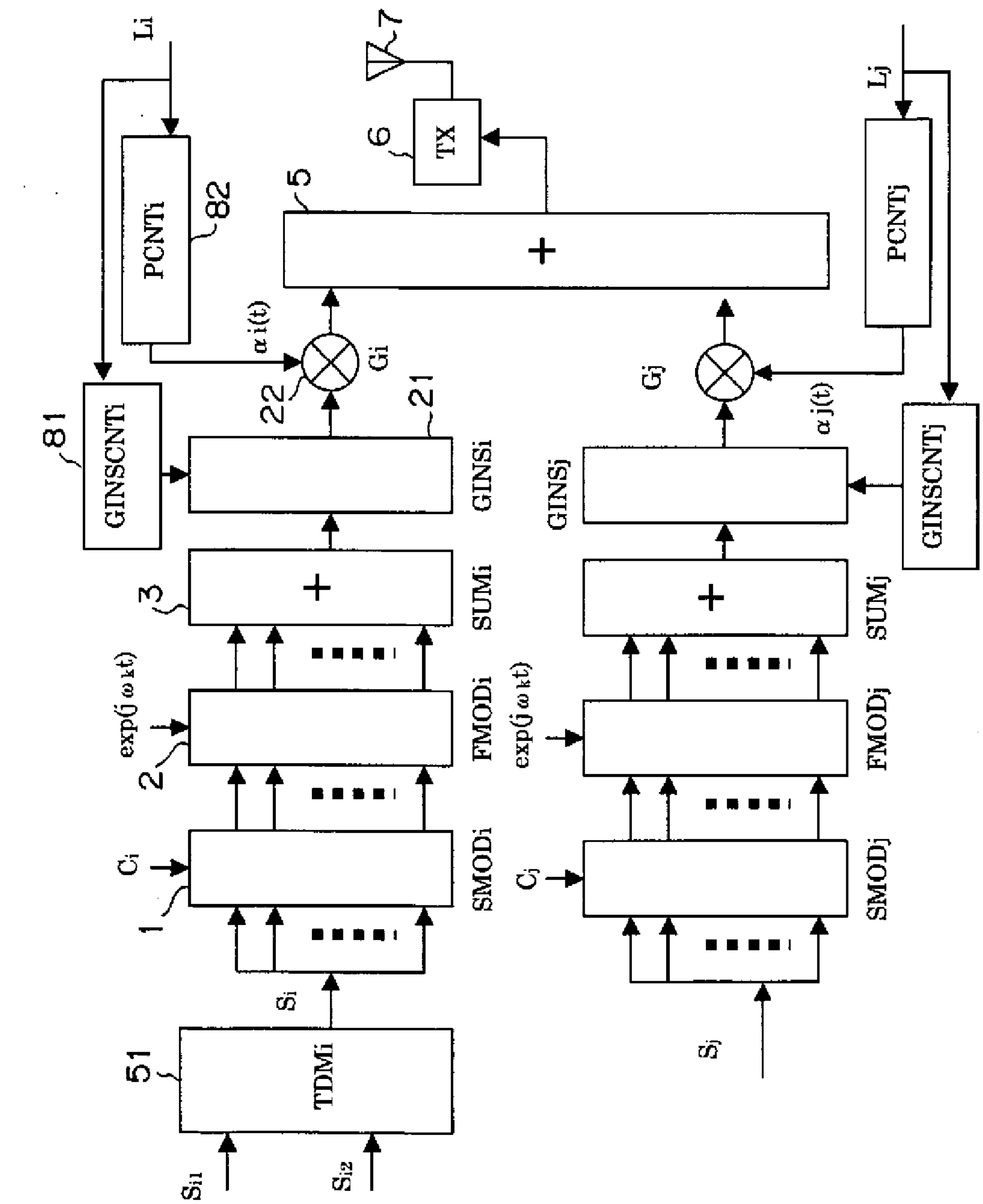

Li
7
TX
6
5
+
PCNTi
82
Lj
PCNTj
αi(t)
Gi
22
21
81
GINSCNTi
GINSi
Gj
αj(t)
GINSj
GINSCNTj
3
+
SUMi
+
SUMj
exp(jωkt)
2
FMODi
exp(jωkt)
FMODj
Ci
1
SMODi
Cj
SMODj
Si
Sj
51
TDMi
Si1
Si2

図26

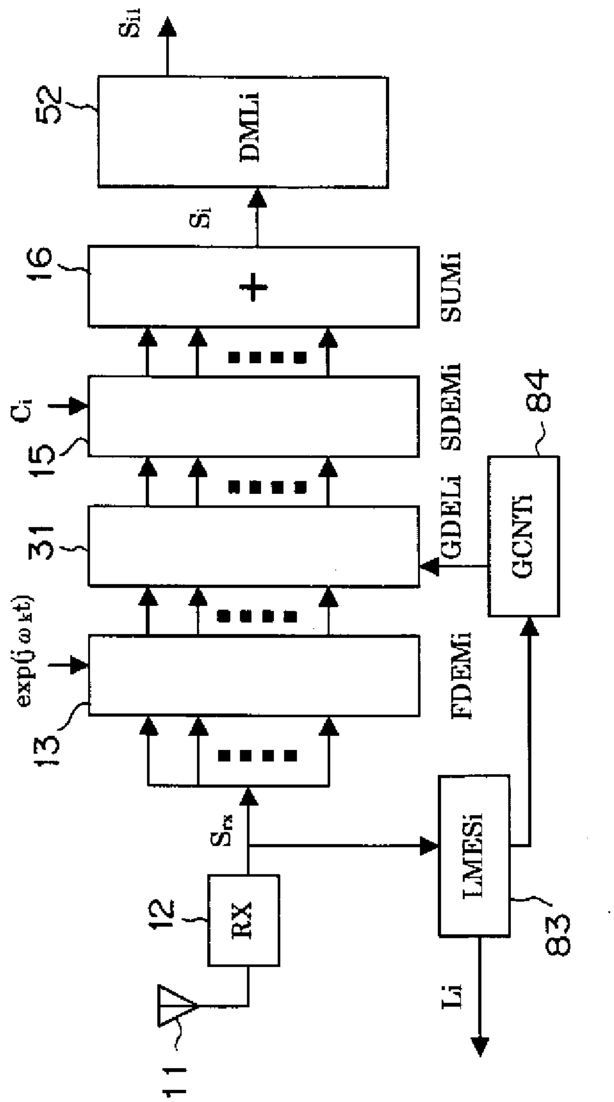

Si1
52
DMLi
Si
16
+
SUMi
Ci
15
SDEMi
84
31
GDELi
GCNTi
exp(jωkt)
13
FDEMi
Srx
12
RX
LMESi
83
Li
11

図27

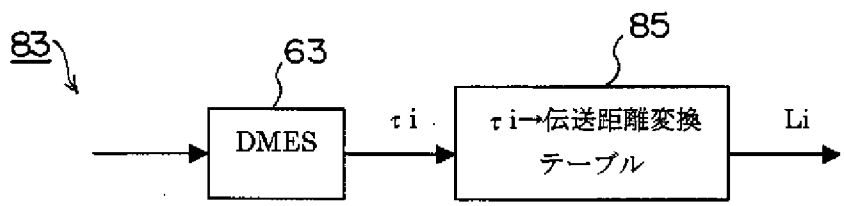


図28

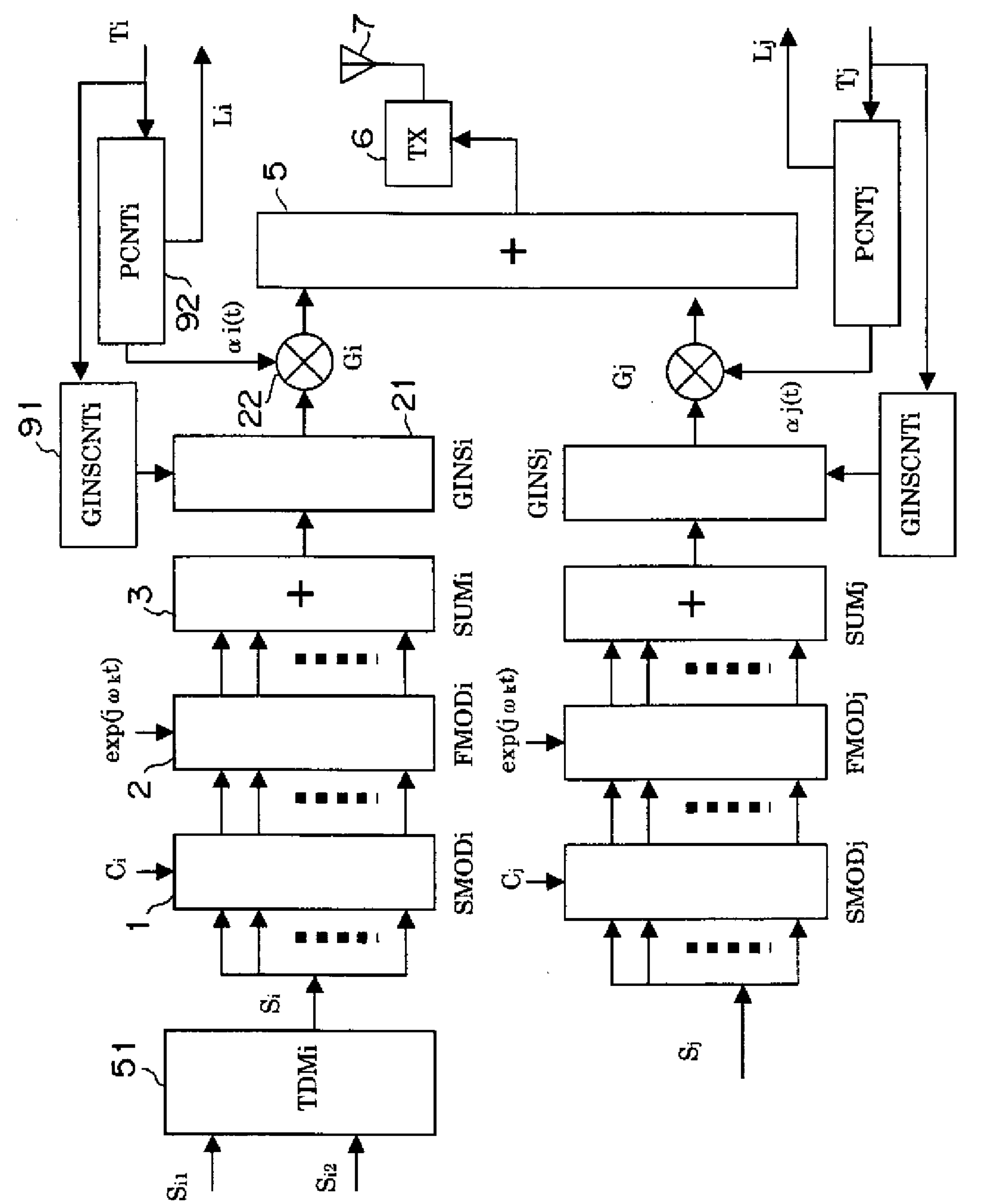

Ti
Li
PCNTi
92
91
GINSCNTi
α i(t)
Gi
22
21
GINSi
3
SUMi
exp(j ω kt)
2
FMODi
Ci
1
SMODi
Si
51
TDMi
Si1
Si2
7
6
TX
5
Lj
Tj
PCNTj
Gj
α j(t)
GINSCNTi
GINSj
SUMj
exp(j ω kt)
FMODj
Cj
SMODj
Sj

図29

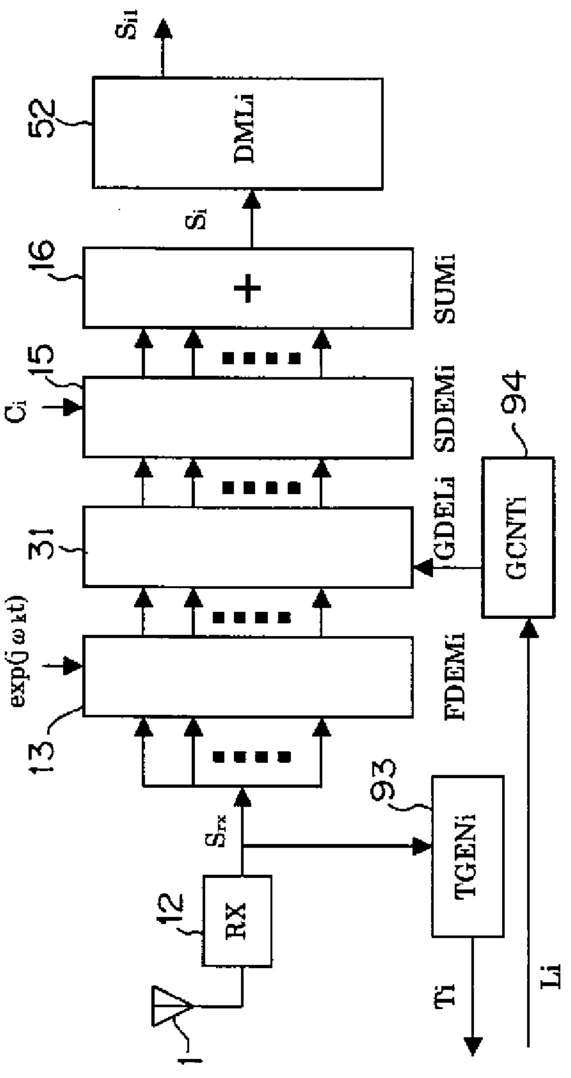

11
12
RX
S_{rx}
13
exp(jω$_k$t)
FDEMi
31
GDELi
15
C_i
SDEMi
16
+
SUMi
S_i
52
DMLi
S_{i1}
93
TGENi
Ti
94
GCNTi
Li

図30

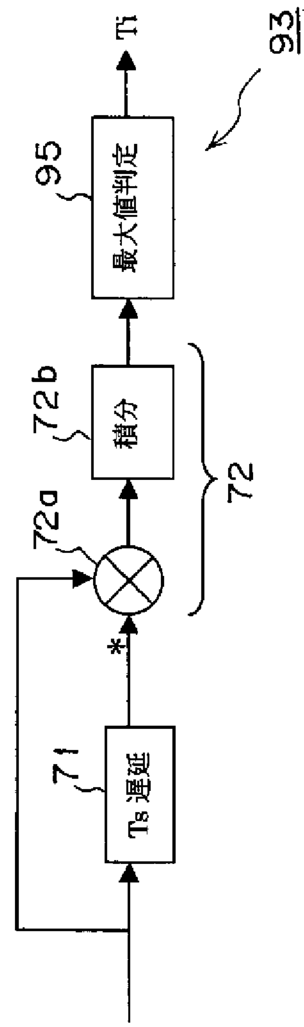


図31

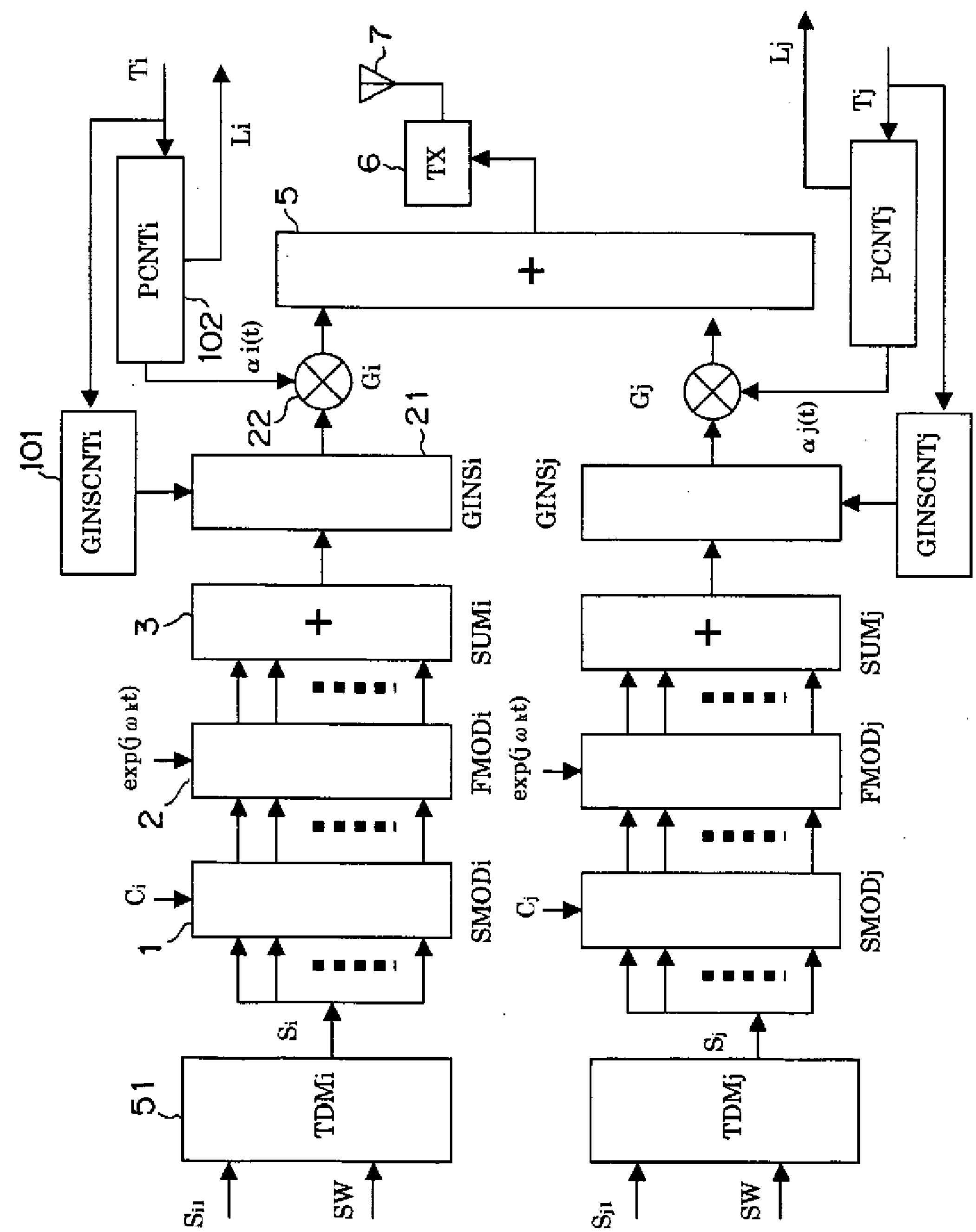

Ti
Li
PCNTi
102
α i(t)
Gi
101
GINSCNTi
22
21
GINSi
3
SUMi
exp(jωkt)
2
FMODi
Ci
1
SMODi
Si
51
TDMi
Si1
SW
7
6
TX
5
Lj
Tj
PCNTj
Gj
α j(t)
GINSCNTj
GINSj
SUMj
exp(jωkt)
FMODj
Cj
SMODj
Sj
TDMj
Sj1
SW

図32

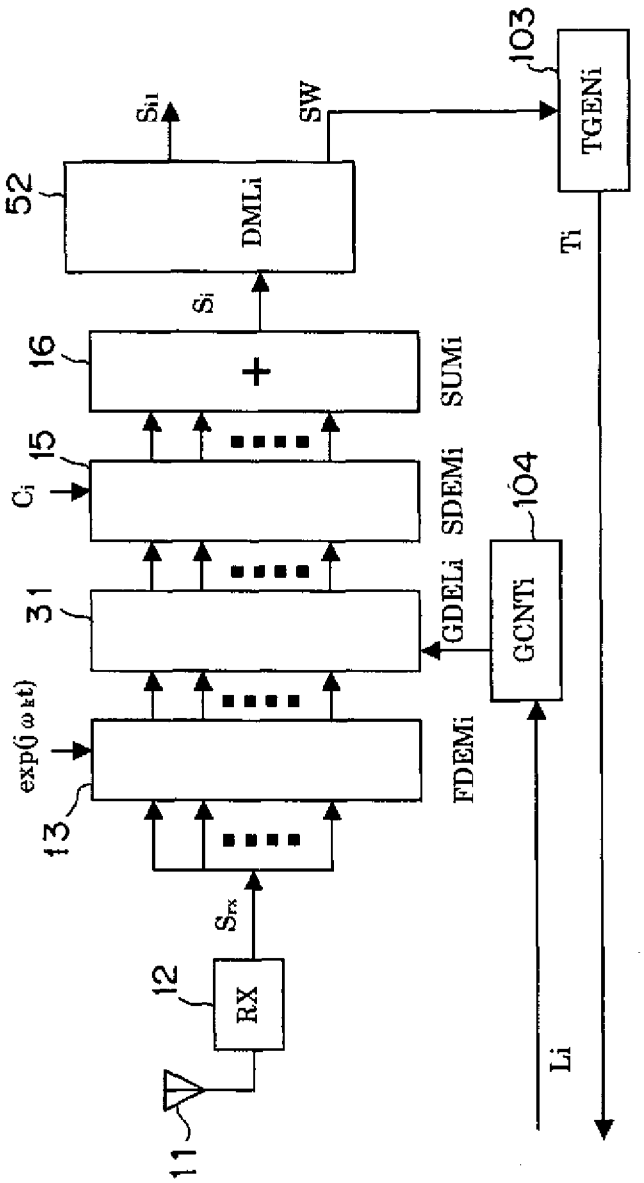

11
12
RX
S_{rx}
13
$\exp(j\omega_k t)$
FDEMi
31
GDELi
GCNTi
104
C_i
15
SDEMi
16
+
SUMi
S_i
52
DMLi
S_{i1}
SW
103
TGENi
Ti
Li

図33

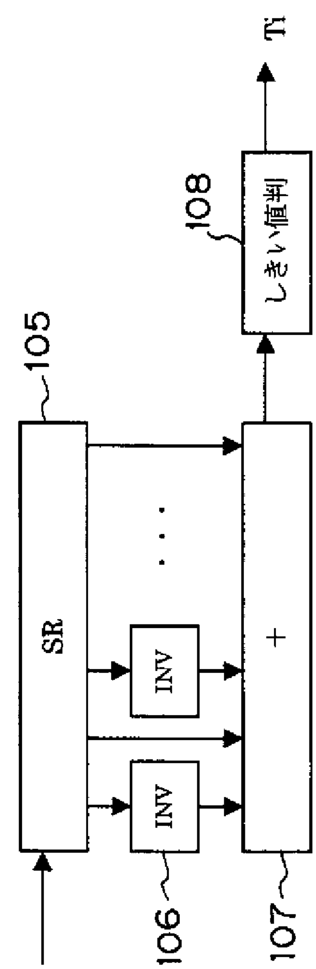


図34

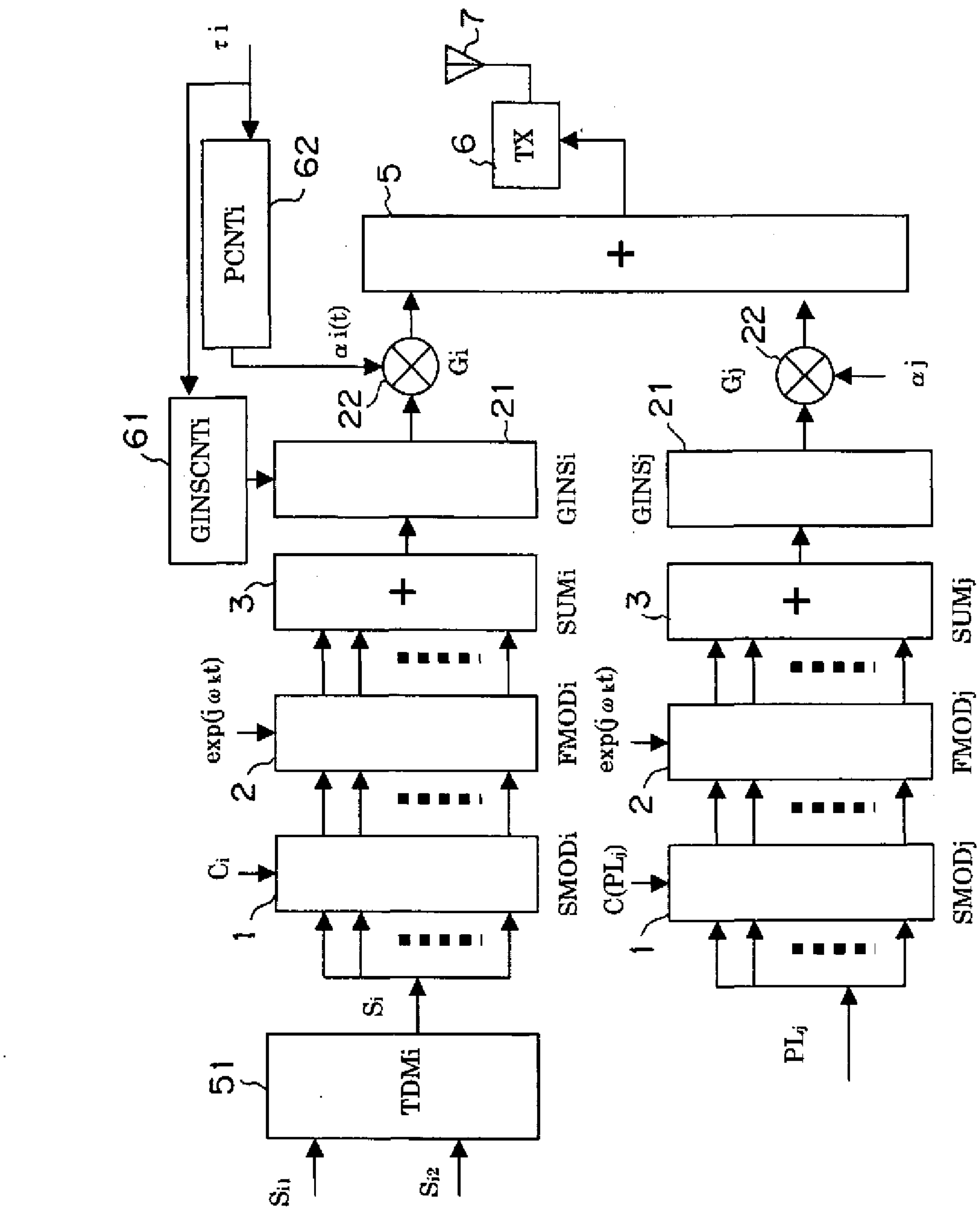

τi
7
PCNTi
62
6
TX
5
+
αi(t)
Gi
22
61
GINSCNTi
21
GINSi
3
+
SUMi
exp(jωkt)
2
FMODi
Ci
1
SMODi
Si
51
TDMi
Si1
Si2
22
Gj
αj
21
GINSj
3
+
SUMj
exp(jωkt)
2
FMODj
C(PLj)
1
SMODj
PLj

図35

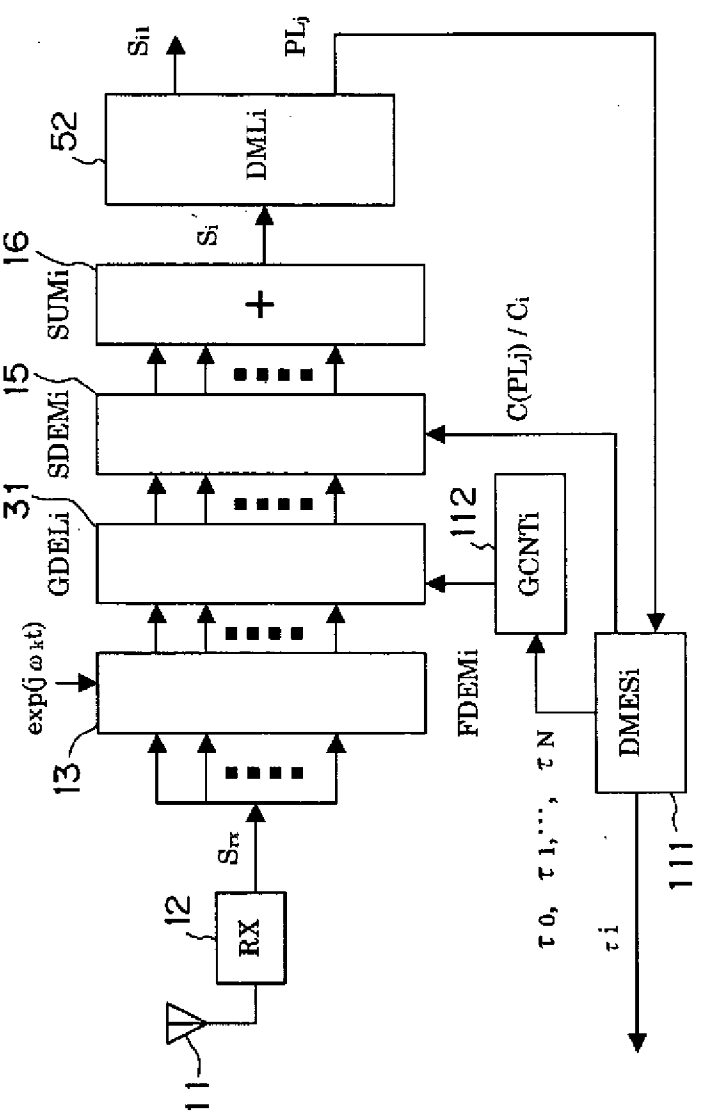
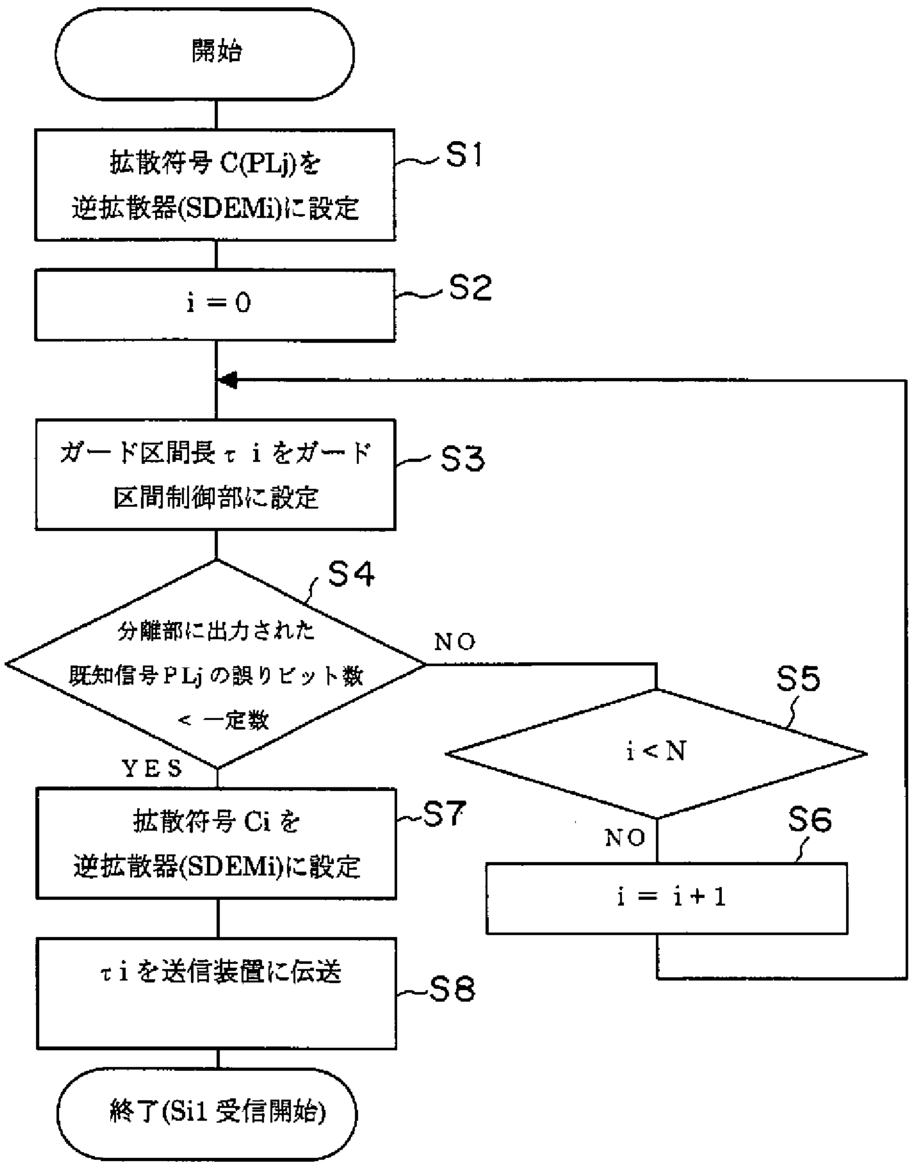

11
12
RX
S_{rx}
13
$\exp(j\omega_k t)$
FDEMi
31
GDELi
112
GCNTi
15
SDEMi
16
SUMi
+
S_i'
52
DMLi
S_{i1}
PL_j
$C(PL_j) / C_i$
111
DMESi
$\tau 0, \tau 1, \cdots, \tau N$
τi

図36



開始
拡散符号 C(PLj)を
逆拡散器(SDEMi)に設定
S1
i＝0
S2
ガード区間長τｉをガード
区間制御部に設定
S3
S4
分離部に出力された
既知信号ＰＬｊの誤りビット数
＜ 一定数
NO
YES
S5
i＜N
NO
S6
i ＝ i＋1
拡散符号 Ci を
逆拡散器(SDEMi)に設定
S7
τｉを送信装置に伝送
S8
終了(Si1 受信開始)

図37

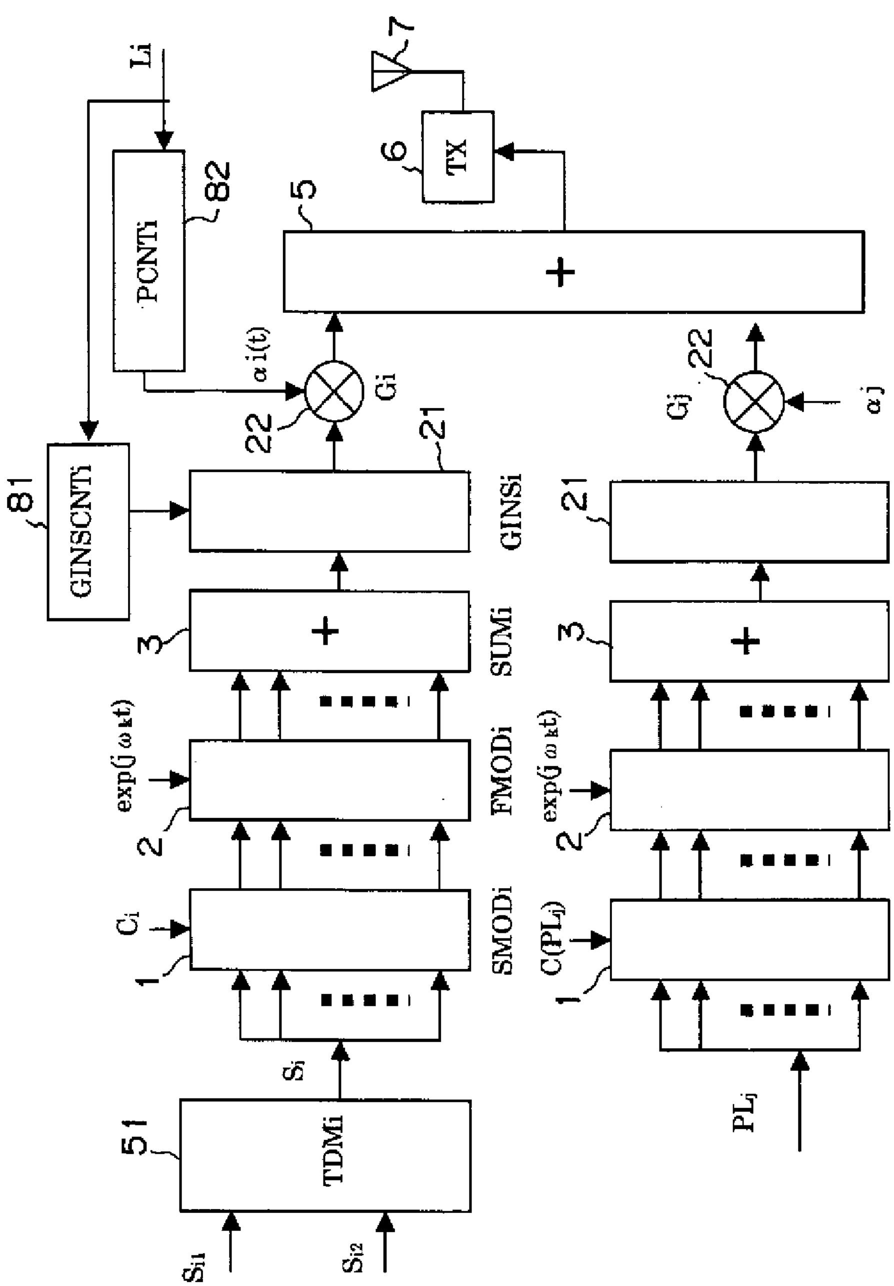

Li
PCNTi
82
GINSCNTi
81
TX
6
7
5
+
αi(t)
Gi
22
21
GINSi
SUMi
3
FMODi
exp(jωkt)
2
SMODi
Ci
1
TDMi
51
Si
Si1
Si2
Gj
αj
GINSj
SUMj
FMODj
SMODj
C(PLj)
PLj

図38

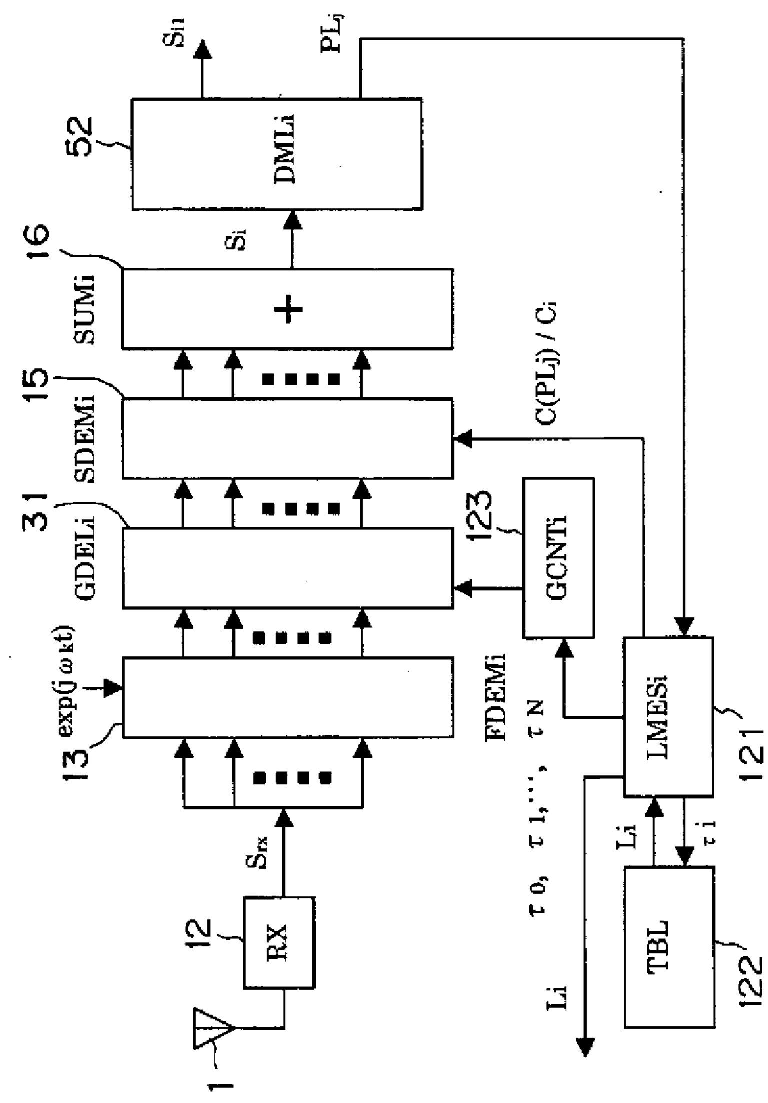

11
12
RX
Srx
13
exp(jωkt)
FDEMi
31
GDELi
123
GCNTi
15
SDEMi
C(PLj) / Ci
16
SUMi
+
Si
52
DMLi
Si1
PLj
121
LMESi
122
TBL
Li
τi
τ0, τ1,…, τN
Li

図39

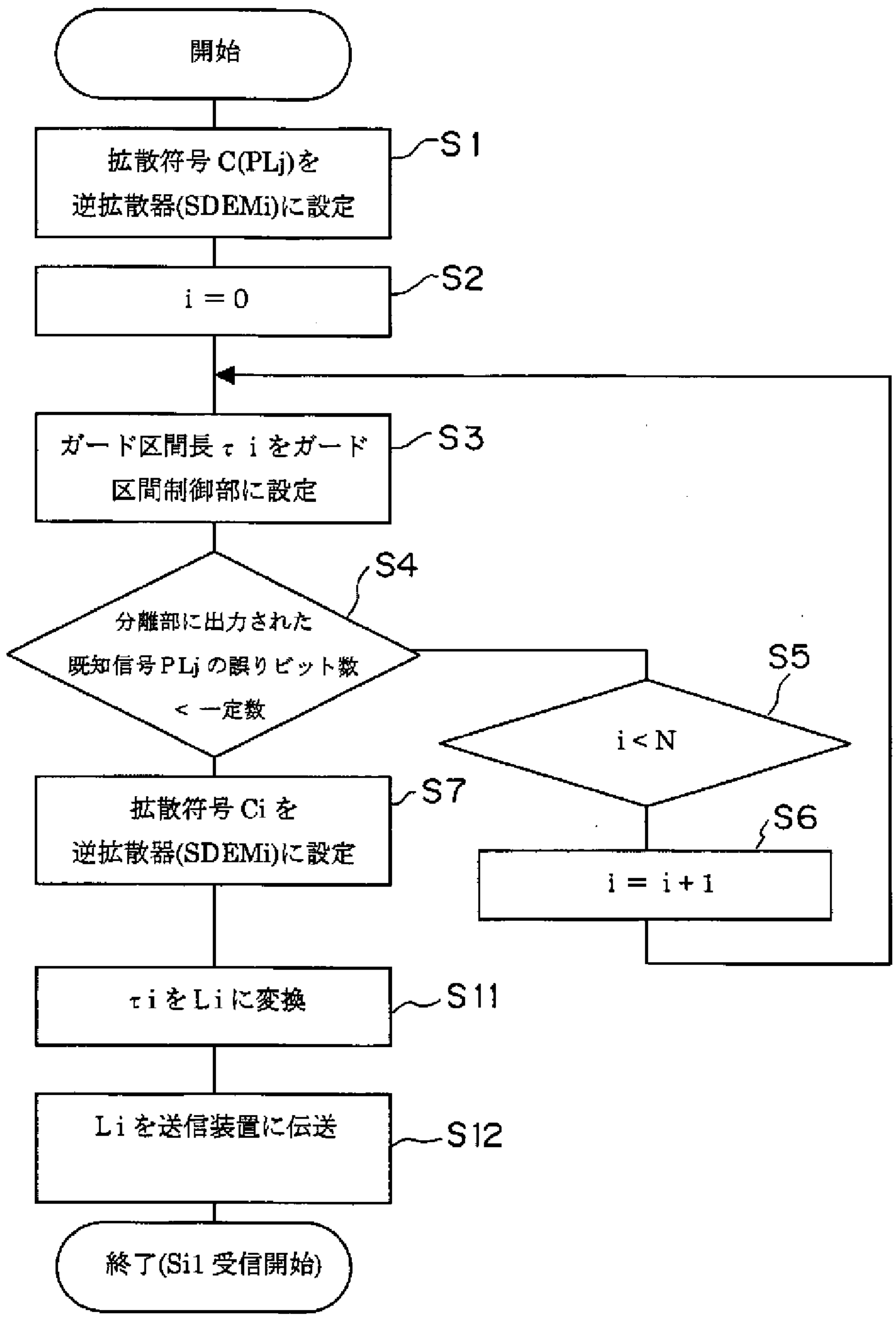

開始
拡散符号C(PLj)を逆拡散器(SDEMi)に設定
S1
i＝0
S2
ガード区間長τiをガード区間制御部に設定
S3
分離部に出力された既知信号PLjの誤りビット数＜一定数
S4
i＜N
S5
i＝i+1
S6
拡散符号Ciを逆拡散器(SDEMi)に設定
S7
τiをLiに変換
S11
Liを送信装置に伝送
S12
終了(Si1受信開始)

図40

INTERNATIONAL SEARCH REPORT

International application No.
PCT/JP01/10357

A. CLASSIFICATION OF SUBJECT MATTER
Int.Cl^7 H04J11/00

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
Int.Cl^7 H04J11/00

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
Jitsuyo Shinan Koho 1926-2000
Kokai Jitsuyo Shinan Koho 1971-2000

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	EP 1035693 A (Matsushita Electric Industrial Co., Ltd.), 13 September 2000 (13.09.2000), page 43, lines 7 to 40; Fig.146 & JP 7-99522 A, page 36, left column, line 35 to right column, line 32	1,3-5,11,13, 16,18-20
A		7-9
X	JP 2000-165342 A (Matsushita Electric Ind. Co., Ltd.), 16 June 2000 (16.06.2000), page 3, left column, line 42 to right column, line 17	1,2,6,10-12, 14-17
A	(Family: none)	7-9
X	EP 1014639 A (Matsushita Electric Industrial Co., Ltd.), 28 June 2000 (28.06.2000), page 4, lines 13 to 18 & JP 2000-244441 A, page 6, right column, lines 22 to 34 & CN 1260649 A & KR 2000052538 A	1,6,10,11, 14-16
A		7-9
A	JP 2001-111519 A (Matsushita Electric Ind. Co., Ltd.), 20 February 2001 (20.04.2001), page 3, right column, lines 9 to 17	1-20

☒ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

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Date of the actual completion of the international search	Date of mailing of the international search report
29 January, 2002 (29.01.02)	05 February, 2002 (05.02.02)
Name and mailing address of the ISA/ Japanese Patent Office	Authorized officer
Facsimile No.	Telephone No.

C (Continuation). DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
	(Family: none)	
A	JP 11-196062 A (Advanced Digital Television Broadcasting), 21 July 1999 (21.07.1991), page 2, right column, lines 37 to 42 (Family: none)	1-20

国際調査報告

国際出願番号　ＰＣＴ／ＪＰ０１／１０３５７

A.　発明の属する分野の分類（国際特許分類（ＩＰＣ））
Ｉｎｔ．Ｃｌ^{7}　Ｈ０４Ｊ１１／００

B.　調査を行った分野
調査を行った最小限資料（国際特許分類（ＩＰＣ））
Ｉｎｔ．Ｃｌ^{7}　Ｈ０４Ｊ１１／００

最小限資料以外の資料で調査を行った分野に含まれるもの
日本国実用新案公報　　１９２６－２０００
日本国公開実用新案公報　　１９７１－２０００

国際調査で使用した電子データベース（データベースの名称、調査に使用した用語）

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X A	ＥＰ　１０３５６９３　Ａ（Ｍａｔｓｕｓｈｉｔａ　Ｅｌｅｃｔｒｉｃ　Ｉｎｄｕｓｔｒｉａｌ　Ｃｏ．，　Ｌｔｄ．），２０００．０９．１３，第４３頁第７行目－第４０行目，ＦＩＧ．１４６ ＆　ＪＰ　７－９９５２２　Ａ，第３６頁左欄第３５行目－右欄第３２行目	1, 3-5, 11, 13, 16, 18-20 7-9

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国際調査を完了した日　２９．０１．０２	国際調査報告の発送日　05.02.02
国際調査機関の名称及びあて先 日本国特許庁（ＩＳＡ／ＪＰ） 郵便番号１００－８９１５ 東京都千代田区霞が関三丁目４番３号	特許庁審査官（権限のある職員）　５Ｋ　９６４７ 高野　洋 電話番号　０３－３５８１－１１０１　内線　３５５５

様式ＰＣＴ／ＩＳＡ／２１０（第２ページ）（１９９８年７月）

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引用文献のカテゴリー＊	引用文献名　及び一部の箇所が関連するときは、その関連する箇所の表示	関連する請求の範囲の番号
X	ＪＰ　２０００－１６５３４２　Ａ（松下電器産業株式会社），２０００．０６．１６，第３頁左欄第４２行目－右欄第１７行目（ファミリーなし）	1, 2, 6, 10-12, 14-17
A		7-9
X	ＥＰ　１０１４６３９　Ａ（Ｍａｔｓｕｓｈｉｔａ　Ｅｌｅｃｔｒｉｃ　Ｉｎｄｕｓｔｒｉａｌ　Ｃｏ．，　Ｌｔｄ．），２０００．０６．２８，第４頁第１３行目－第１８行目 ＆　ＪＰ　２０００－２４４４４１　Ａ，第６頁右欄第２２行目－第３４行目 ＆　ＣＮ　１２６０６４９　Ａ ＆　ＫＲ　２０００００５２５３８　Ａ	1, 6, 10, 11, 14-16
A		7-9
A	ＪＰ　２００１－１１１５１９　Ａ（松下電器産業株式会社），２００１．０４．２０，第３頁右欄第９行目－第１７行目（ファミリーなし）	1-20
A	ＪＰ　１１－１９６０６２　Ａ（株式会社次世代デジタルテレビジョン放送システム研究所），１９９９．０７．２１，第２頁右欄第３７行目－第４２行目（ファミリーなし）	1-20

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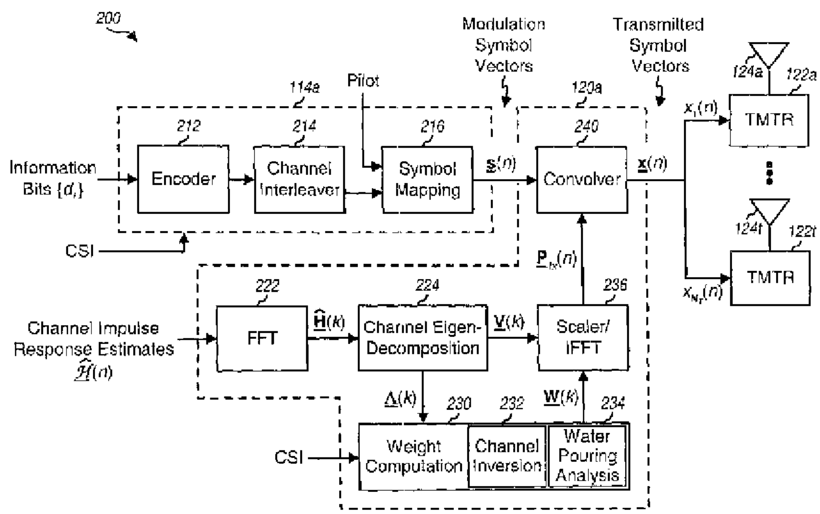
(54) **Title:** SIGNAL PROCESSING WITH CHANNEL EIGENMODE DECOMPOSITION AND CHANNEL INVERSION FOR MIMO SYSTEMS



(57) **Abstract:** Techniques for processing a data transmission at a transmitter and receiver, which use channel eigen-decomposition, channel inversion, and (optionally) "water-pouring". At the transmitter, (1) channel eigen-decomposition is performed to determine eigenmodes of a MIMO channel and to derive a first set of steering vectors, (2) channel inversion is performed to derive weights (*e.g.*, one set for each eigenmode) used to minimize ISI distortion, and (3) water-pouring may be performed to derive scaling values indicative of the transmit powers allocated to the eigenmodes. The first set of steering vectors, weights, and scaling values are used to derive a pulse-shaping matrix, which is used to precondition modulation symbols prior to transmission. At the receiver, channel eigen-decomposition is performed to derive a second set of steering vectors, which are used to derive a pulse-shaping matrix used to condition received symbols such that orthogonal symbol streams are recovered.

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SIGNAL PROCESSING WITH CHANNEL EIGENMODE DECOMPOSITION AND CHANNEL INVERSION FOR MIMO SYSTEMS

BACKGROUND

Field

[1001] The present invention relates generally to data communication, and more specifically to techniques for performing signal processing with channel eigenmode decomposition and channel inversion for multiple-input multiple-output (MIMO) communication systems.

Background

[1002] A multiple-input multiple-output (MIMO) communication system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels is also referred to as a spatial subchannel of the MIMO channel and corresponds to a dimension. The MIMO system can provide improved performance (e.g., increased transmission capacity) if the additional dimensionalities created by the multiple transmit and receive antennas are utilized.

[1003] The spatial subchannels of a wideband MIMO system may encounter different channel conditions due to various factors such as fading and multipath. Each spatial subchannel may thus experience frequency selective fading, which is characterized by different channel gains at different frequencies (i.e., different frequency bins or subbands) of the overall system bandwidth. With frequency selective fading, each spatial subchannel may achieve different signal-to-noise-and-interference ratios (SNRs) for different frequency bins. Consequently, the number of information bits per modulation symbol (or data rate) that may be transmitted at different frequency bins of each spatial subchannel for a particular level of performance (e.g., 1% packet error rate) may be different from bin to bin. Moreover, because the channel conditions

typically vary with time, the supported data rates for the bins of the spatial subchannels also vary with time.

[1004] To combat frequency selective fading in a wideband channel, orthogonal frequency division multiplexing (OFDM) may be used to effectively partition the system bandwidth into a number of (N_F) subbands (which may also be referred to as frequency bins or subchannels). With OFDM, each frequency subchannel is associated with a respective subcarrier upon which data may be modulated. For a MIMO system that utilizes OFDM (i.e., a MIMO-OFDM system), each frequency subchannel of each spatial subchannel may be viewed as an independent transmission channel.

[1005] A key challenge in a coded communication system is the selection of the appropriate data rates and coding and modulation schemes to be used for a data transmission based on channel conditions. The goal of this selection process is to maximize throughput while meeting quality objectives, which may be quantified by a particular packet error rate (PER), certain latency criteria, and so on.

[1006] One straightforward technique for selecting data rates and coding and modulation schemes is to "bit load" each transmission channel in the MIMO-OFDM system according to its transmission capability, which may be quantified by the channel's short-term average SNR. However, this technique has several major drawbacks. First, coding and modulating individually for each transmission channel can significantly increase the complexity of the processing at both the transmitter and receiver. Second, coding individually for each transmission channel may greatly increase coding and decoding delay. And third, a high feedback rate would be needed to send channel state information (CSI) indicative of the channel conditions (e.g., the gain, phase, and SNR) of each transmission channel.

[1007] For a MIMO system, transmit power is another parameter that may be manipulated to maximize throughput. In general, the overall throughput of the MIMO system may be increased my allocating more transmit power to transmission channels with greater transmission capabilities. However, allocating different amounts of transmit power to different frequency bins of a given spatial subchannel tends to exaggerate the frequency selective nature of the spatial subchannel. It is well known that frequency selective fading causes inter-symbol interference (ISI), which is a phenomenon whereby each symbol in a received signal acts as distortion to subsequent symbols in the received signal. The ISI distortion degrades performance by impacting

the ability to correctly detect the received symbols. To mitigate the deleterious effects of ISI, equalization of the received symbols would need to be performed at the receiver. Thus, a major drawback in frequency-domain power allocation is the additional complexity at the receiver to combat the resultant additional ISI distortion.

[1008] There is therefore a need in the art for techniques to achieve high overall throughput in a MIMO system without having to individually code each transmission channel and which mitigate the deleterious effects of ISI.

SUMMARY

[1009] Techniques are provided herein for processing a data transmission at a transmitter and a receiver of a MIMO system such that high performance (e.g., high overall throughput) is achieved. In an aspect, a time-domain implementation is provided which uses frequency-domain channel eigen-decomposition, channel inversion, and (optionally) "water-pouring" results to derive pulse-shaping and beam-steering solutions for the transmitter and receiver.

[1010] Channel eigen-decomposition is performed at the transmitter to determine the eigenmodes (i.e., the spatial subchannels) of a MIMO channel and to obtain a first set of steering vectors, which are used to precondition modulation symbols prior to transmission over the MIMO channel. Channel eigen-decomposition may be performed based on an estimated channel response matrix, which is an estimate of the (time-domain or frequency-domain) channel response of the MIMO channel. Channel eigen-decomposition is also performed at the receiver to obtain a second set of steering vectors, which are used to condition received symbols such that orthogonal symbol streams are recovered at the receiver.

[1011] Channel inversion is performed at the transmitter to derive weights, which are used to minimize or reduce the amount of ISI distortion at the receiver. In particular, the channel inversion may be performed for each eigenmode used for data transmission. One set of weights may be derived for each eigenmode based on the estimated channel response matrix for the MIMO channel and is used to invert the frequency response of the eigenmode.

[1012] Water-pouring analysis may (optionally) be used to more optimally allocate the total available transmit power to the eigenmodes of the MIMO channel. In particular, eigenmodes with greater transmission capabilities may be allocated more

transmit power, and eigenmodes with transmission capabilities below a particular threshold may be omitted from use (e.g., by allocating these bad eigenmodes with zero transmit power). The transmit power allocated to each eigenmode then determines the data rate and possibly the coding and modulation scheme to be used for the eigenmode.

[1013] At the transmitter, data is initially processed (e.g., coded and modulated) in accordance with a particular processing scheme to provide a number of modulation symbol streams (e.g., one modulation symbol stream for each eigenmode). An estimated channel response matrix for the MIMO channel is obtained (e.g., from the receiver) and decomposed (e.g., in the frequency domain, using channel eigen-decomposition) to obtain a set of matrices of right eigen-vectors and a set of matrices of singular values. A number of sets of weights are then derived based on the matrices of singular values, with each set of weights being used to invert the frequency response of a respective eigenmode used for data transmission. Water-pouring analysis may also be performed based on the matrices of singular values to obtain a set of scaling values indicative of the transmit powers allocated to the eigenmodes. A pulse-shaping matrix for the transmitter is then derived based on the matrices of right eigen-vectors, the weights, and the scaling values (if available). The pulse-shaping matrix comprises steering vectors, which are used to precondition the streams of modulation symbols to obtain streams of preconditioned symbols to be transmitted over the MIMO channel.

[1014] At the receiver, the estimated channel response matrix is also obtained (e.g., based on pilot symbols sent from the transmitter) and decomposed to obtain a set of matrices of left eigen-vectors. A pulse-shaping matrix for the receiver is then derived based on the matrices of left eigen-vectors and used to condition a number of received symbol streams to obtain a number of recovered symbol streams. The recovered symbols are further processed (e.g., demodulated and decoded) to recover the transmitted data.

[1015] Various aspects and embodiments of the invention are described in further detail below. The invention further provides methods, digital signal processors, transmitter and receiver units, and other apparatuses and elements that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1016] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1017] FIG. 1 is a block diagram of an embodiment of a transmitter system and a receiver system in a MIMO system;

[1018] FIG. 2 is a block diagram of a transmitter unit within the transmitter system;

[1019] FIGS. 3A and 3B are diagrams that graphically illustrate the derivation of the weights used to invert the frequency response of each eigenmode of a MIMO channel;

[1020] FIG. 4 is a flow diagram of a process for allocating the total available transmit power to the eigenmodes of the MIMO channel;

[1021] FIGS. 5A and 5B are diagrams that graphically illustrate the allocation of the total transmit power to three eigenmodes in an example MIMO system;

[1022] FIG. 6 is a flow diagram of an embodiment of the signal processing at the transmitter unit;

[1023] FIG. 7 is a block diagram of a receiver unit within the receiver system; and

[1024] FIG. 8 is a flow diagram of an embodiment of the signal processing at the receiver unit.

DETAILED DESCRIPTION

[1025] The techniques described herein for processing a data transmission at a transmitter and receiver may be used for various wireless communication systems. For clarity, various aspects and embodiments of the invention are described specifically for a multiple-input multiple-output (MIMO) communication system.

[1026] A MIMO system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels is also referred to as a spatial subchannel of the MIMO channel. The number of spatial subchannels is determined by the number of eigenmodes for the MIMO channel, which in turn is dependent on a

channel response matrix that describes the response between the N_T transmit and N_R receive antennas.

[1027] **FIG. 1** is a block diagram of an embodiment of a transmitter system 110 and a receiver system 150, which are capable of implementing various signal processing techniques described herein.

[1028] At transmitter system 110, traffic data is provided from a data source 112 to a transmit (TX) data processor 114, which formats, codes, and interleaves the traffic data based on one or more coding schemes to provide coded data. The coded traffic data may then be multiplexed with pilot data using, for example, time division multiplex (TDM) or code division multiplex (CDM), in all or a subset of the data streams to be transmitted. The pilot data is typically a known data pattern processed in a known manner, if at all. The multiplexed pilot and coded traffic data is interleaved and then modulated (i.e., symbol mapped) based on one or more modulation schemes to provide modulation symbols. In an embodiment, TX data processor 114 provides one modulation symbol stream for each spatial subchannel used for data transmission. The data rate, coding, interleaving, and modulation for each modulation symbol stream may be determined by controls provided by a controller 130.

[1029] The modulation symbols are then provided to a TX MIMO processor 120 and further processed. In a specific embodiment, the processing by TX MIMO processor 120 includes (1) determining an estimated channel frequency response matrix for the MIMO channel, (2) decomposing this matrix to determine the eigenmodes of the MIMO channel and to derive a set of "steering" vectors for the transmitter, one vector for the modulation symbol stream to be transmitted on each spatial subchannel, (3) deriving a transmit spatio-temporal pulse-shaping matrix based on the steering vectors and a weighting matrix indicative of the transmit powers assigned to the frequency bins of the eigenmodes, and (4) preconditioning the modulation symbols with the pulse-shaping matrix to provide preconditioned modulation symbols. The processing by TX MIMO processor 120 is described in further detail below. Up to N_T streams of preconditioned symbols are then provided to transmitters (TMTR) 122a through 122t.

[1030] Each transmitter 122 converts a respective preconditioned symbol stream into one or more analog signals and further conditions (e.g., amplifies, filters, and frequency upconverts) the analog signals to generate a modulated signal suitable for

transmission over the MIMO channel. The modulated signal from each transmitter 122 is then transmitted via a respective antenna 124 to the receiver system.

[1031] At receiver system 150, the transmitted modulated signals are received by N_R antennas 152a through 152r, and the received signal from each antenna 152 is provided to a respective receiver (RCVR) 154. Each receiver 154 conditions (e.g., filters, amplifies, and frequency downconverts) the received signal, digitizes the conditioned signal to provide a stream of samples, and further processes the sample stream to provide a stream of received symbols. An RX MIMO processor 160 then receives and processes the N_R received symbol streams to provide N_T streams of recovered symbols, which are estimates of the modulation symbols transmitted from the transmitter system. In an embodiment, the processing by RX MIMO processor 160 may include (1) determining the estimated channel frequency response matrix for the MIMO channel, (2) decomposing this matrix to derive a set of steering vectors for the receiver, (3) deriving a receive spatio-temporal pulse-shaping matrix based on the steering vectors, and (4) conditioning the received symbols with the pulse-shaping matrix to provide the recovered symbols. The processing by RX MIMO processor 160 is described in further detail below.

[1032] A receive (RX) data processor 162 then demodulates, deinterleaves, and decodes the recovered symbols to provide decoded data, which is an estimate of the transmitted traffic data. The processing by RX MIMO processor 160 and RX data processor 162 is complementary to that performed by TX MIMO processor 120 and TX data processor 114, respectively, at transmitter system 110.

[1033] RX MIMO processor 160 may further derive channel impulse responses for the MIMO channel, received noise power and/or signal-to-noise-and-interference ratios (SNRs) for the spatial subchannels, and so on. RX MIMO processor 160 would then provide these quantities to a controller 170. RX data processor 162 may also provide the status of each received packet or frame, one or more other performance metrics indicative of the decoded results, and possibly other information. Controller 170 then derives channel state information (CSI), which may comprise all or some of the information received from RX MIMO processor 160 and RX data processor 162. The CSI is processed by a TX data processor 178, modulated by a modulator 180, conditioned by transmitters 154a through 154r, and sent back to transmitter system 110.

[1034] At transmitter system 110, the modulated signals from receiver system 150 are received by antennas 124, conditioned by receivers 122, and demodulated by a demodulator 140 to recover the CSI transmitted by the receiver system. The CSI is then provided to controller 130 and used to generate various controls for TX data processor 114 and TX MIMO processor 120.

[1035] Controllers 130 and 170 direct the operation at the transmitter and receiver systems, respectively. Memories 132 and 172 provide storage for program codes and data used by controllers 130 and 170, respectively.

[1036] Techniques are provided herein for achieving high performance (e.g., high overall system throughput) via a time-domain implementation that uses frequency-domain channel eigen-decomposition, channel inversion, and (optionally) water-pouring results to derive time-domain pulse-shaping and beam-steering solutions for the transmitter and receiver.

[1037] Channel eigen-decomposition is performed at the transmitter to determine the eigenmodes of the MIMO channel and to derive a first set of steering vectors, which are used to precondition the modulation symbols. Channel eigen-decomposition is also performed at the receiver to derive a second set of steering vectors, which are used to condition the received symbols such that orthogonal symbol streams are recovered at the receiver. The preconditioning at the transmitter and the conditioning at the receiver orthogonalize the symbol streams transmitted over the MIMO channel.

[1038] Channel inversion is performed at the transmitter to flatten the frequency response of each eigenmode (or spatial subchannel) used for data transmission. As noted above, frequency selective fading causes intersymbol interference (ISI), which can degrade performance by impacting the ability to correctly detect the received symbols at the receiver. Conventionally, the frequency selective fading may be compensated for at the receiver by performing equalization on the received symbol streams. However, equalization increases the complexity of the receiver processing. With the inventive techniques, the channel inversion is performed at the transmitter to account for the frequency selective fading and to mitigate the need for equalization at the receiver.

[1039] Water-pouring (or water-filling) analysis is used to more optimally allocate the total available transmit power in the MIMO system to the eigenmodes such that high performance is achieved. The transmit power allocated to each eigenmode may then

determine the data rate and the coding and modulation scheme to be used for the eigenmode.

[1040] These various processing techniques are described in further detail below.

[1041] The techniques described herein provide several potential advantages. First, with time-domain eigenmode decomposition, the maximum number of eigenmodes with different SNRs is given by $\min(N_T, N_R)$. If one independent data stream is transmitted on each eigenmode and each data stream is independently processed, then the maximum number of different coding/modulation schemes is also given by $\min(N_T, N_R)$. It is also possible to make the received SNRs for the data streams essentially the same, thereby further simplifying the coding/modulation. The techniques described herein can thus greatly simplify the coding/modulation for a data transmission by avoiding the per-bin bit allocation required to approach channel capacity in MIMO-OFDM systems that utilize frequency-domain water-pouring.

[1042] Second, the channel inversion at the transmitter results in recovered symbol streams at the receiver that do not require equalization. This then reduces the complexity of the receiver processing. In contrast, other wide-band time-domain techniques typically require complicated space-time equalization to recover the symbol streams.

[1043] Third, the time-domain signaling techniques described herein can more easily integrate the channel/pilot structures of various CDMA standards, which are also based on time-domain signaling. Implementation of the channel/pilot structures may be more complicated in OFDM-based systems that perform frequency-domain signaling.

[1044] **FIG. 2** is a block diagram of an embodiment of a transmitter unit 200, which is capable of implementing various processing techniques described herein. Transmitter unit 200 is an embodiment of the transmitter portion of transmitter system 110 in FIG. 1. Transmitter unit 200 includes (1) a TX data processor 114a that receives and processes traffic and pilot data to provide N_T modulation symbol streams and (2) a TX MIMO processor 120a that preconditions the modulation symbol streams to provide N_T preconditioned symbol streams. TX data processor 114a and TX MIMO processor 120a are one embodiment of TX data processor 114 and TX MIMO processor 120, respectively, in FIG. 1.

[1045] In the specific embodiment shown in FIG. 2, TX data processor 114a includes an encoder 212, a channel interleaver 214, and a symbol mapping element 216. Encoder 212 receives and codes the traffic data (i.e., the information bits, d_i) in accordance with one or more coding schemes to provide coded bits. The coding increases the reliability of the data transmission. In an embodiment, a separate coding scheme may be used for the information bits for each eigenmode (or spatial subchannel) selected for use for data transmission. In alternative embodiments, a separate coding scheme may be used for each subset of spatial subchannels, or a common coding scheme may be used for all spatial subchannels. The coding scheme(s) to be used are determined by controls from controller 130 and may be selected based on the CSI received from the receiver system. Each selected coding scheme may include any combination of cyclic redundancy check (CRC), convolutional coding, Turbo coding, block coding, and other coding, or no coding at all.

[1046] Channel interleaver 214 interleaves the coded bits based on one or more interleaving schemes. Typically, each selected coding scheme is associated with a corresponding interleaving scheme. The interleaving provides time diversity for the coded bits, permits the data to be transmitted based on an average SNR of each spatial subchannel used for the data transmission, combats fading, and further removes correlation between coded bits used to form each modulation symbol.

[1047] Symbol mapping element 216 then receives and multiplexes pilot data with the interleaved data and further maps the multiplexed data in accordance with one or more modulation schemes to provide modulation symbols. A separate modulation scheme may be used for each spatial subchannel selected for use, or for each subset of spatial subchannels. Alternatively, a common modulation scheme may be used for all selected spatial subchannels.

[1048] The symbol mapping for each spatial subchannel may be achieved by grouping sets of bits to form data symbols (each of which may be a non-binary value) and mapping each data symbol to a point in a signal constellation corresponding to the modulation scheme selected for use for that spatial subchannel. The selected modulation scheme may be QPSK, M-PSK, M-QAM, or some other scheme. Each mapped signal point is a complex value and corresponds to a modulation symbol. Symbol mapping element 216 provides a vector of modulation symbols for each symbol

period, with the number of modulation symbols in each vector corresponding to the number of spatial subchannels selected for use for that symbol period. Symbol mapping element 216 thus provides up to N_T modulation symbol streams. These streams collectively form a sequence of vectors, with are also referred to as the modulation symbol vectors, $\underline{s}(n)$, with each such vector including up to N_S modulation symbols to be transmitted on up to N_S spatial subchannels for the n-th symbol period.

[1049] Within TX MIMO processor 120a, the response of the MIMO channel is estimated and used to precondition the modulation symbols prior to transmission to the receiver system. In a frequency division duplexed (FDD) system, the downlink and uplink are allocated different frequency bands, and the channel responses for the downlink and uplink may not be correlated to a sufficient degree. For the FDD system, the channel response may be estimated at the receiver and sent back to the transmitter. In a time division duplexed (TDD) system, the downlink and uplink share the same frequency band in a time division multiplexed manner, and a high degree of correlation may exist between the downlink and uplink channel responses. For the TDD system, the transmitter system may estimate the uplink channel response (e.g., based on the pilot transmitted by the receiver system on the uplink) and may then derive the downlink channel response by accounting for any differences such as those between the transmit and receive antenna array manifolds.

[1050] In an embodiment, the channel response estimates are provided to TX MIMO processor 120a as a sequence of $N_R \times N_T$ matrices, $\underline{\hat{\mathcal{H}}}(n)$, of time-domain samples. This sequence of matrices is collectively referred to as a channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$. The (i, j)-th element, $\underline{\hat{h}}_{i,j}$, of the estimated channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$, for $i = (1, 2, \ldots, N_R)$ and $j = (1, 2, \ldots, N_T)$, is a sequence of samples that represents the sampled impulse response of the propagation path from the j-th transmit antenna to the i-th receive antenna.

[1051] Within TX MIMO processor 120a, a fast Fourier transformer 222 receives the estimated channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$ (e.g., from the receiver system) and derives the corresponding estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$, by performing a fast Fourier transform (FFT) on the matrix $\underline{\underline{\hat{\mathcal{H}}}}$ (i.e., $\underline{\underline{\hat{\mathbf{H}}}} = \text{FFT}[\underline{\underline{\hat{\mathcal{H}}}}]$). This

may be achieved by performing an N_F-point FFT on a sequence of N_F samples for each element of $\underline{\underline{\hat{\mathcal{H}}}}$ to derive a set of N_F coefficients for the corresponding element of $\underline{\underline{\hat{\mathbf{H}}}}$, where N_F corresponds to the number of frequency bins for the FFT (i.e., the length of the FFT). The $N_R \cdot N_T$ elements of $\underline{\underline{\hat{\mathbf{H}}}}$ are thus $N_R \cdot N_T$ sets of coefficients representing the frequency responses of the propagation paths between the N_T transmit antennas and N_R receive antennas. Each element of $\underline{\underline{\hat{\mathbf{H}}}}$ is the FFT of the corresponding element of $\underline{\underline{\hat{\mathcal{H}}}}$. The estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$, may also be viewed as comprising a set of N_F matrices, $\underline{\hat{\mathbf{H}}}(k)$ for $k = (0,\ 1,\ \ldots\ , N_F - 1)$.

Channel Eigen-Decomposition

[1052] A unit 224 then performs eigen-decomposition of the MIMO channel used for data transmission. In one embodiment for performing channel eigen-decomposition, unit 224 computes the singular value decomposition (SVD) of the estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$. In an embodiment, the singular value decomposition is performed for each matrix $\underline{\hat{\mathbf{H}}}(k)$, for $k = (0,\ 1,\ \ldots\ , N_F - 1)$. The singular value decomposition of matrix $\underline{\hat{\mathbf{H}}}(k)$ for frequency bin k (or frequency f_k) may be expressed as:

$$\underline{\hat{\mathbf{H}}}(k) = \underline{\mathbf{U}}(k)\underline{\Lambda}(k)\underline{\mathbf{V}}^H(k) \ , \qquad \text{Eq (1)}$$

where $\underline{\mathbf{U}}(k)$ is an $N_R \times N_R$ unitary matrix (i.e., $\underline{\mathbf{U}}^H\underline{\mathbf{U}} = \underline{\mathbf{I}}$, where $\underline{\mathbf{I}}$ is the identity matrix with ones along the diagonal and zeros everywhere else);

$\underline{\Lambda}(k)$ is an $N_R \times N_T$ diagonal matrix of singular values of $\underline{\hat{\mathbf{H}}}(k)$; and

$\underline{\mathbf{V}}(k)$ is an $N_T \times N_T$ unitary matrix.

The diagonal matrix $\underline{\Lambda}(k)$ contains non-negative real values along the diagonal (i.e., $\underline{\Lambda}(k) = \text{diag}\,(\lambda_1(k),\ \lambda_2(k),\ \ldots\ , \lambda_{N_T}(k))$) and zeros elsewhere. The $\lambda_i(k)$, for $i = (1,\ 2,\ \ldots\ , N_T)$, are referred to as the singular values of the matrix $\underline{\hat{\mathbf{H}}}(k)$. The singular value decomposition is a matrix operation known in the art and described in various references. One such reference is a book by Gilbert Strang entitled "Linear

Algebra and Its Applications," Second Edition, Academic Press, 1980, which is incorporated herein by reference.

[1053] The result of the singular value decomposition is three sets of N_F matrices, $\underline{\underline{\mathbf{U}}}$, $\underline{\underline{\Lambda}}$, and $\underline{\underline{\mathbf{V}}}^H$, where $\underline{\underline{\mathbf{U}}} = [\underline{\mathbf{U}}(0) \ \dots \ \underline{\mathbf{U}}(k) \ \dots \ \underline{\mathbf{U}}(N_F - 1)]$, and so on. For each value of k, $\underline{\mathbf{U}}(k)$ is the $N_R \times N_R$ unitary matrix of left eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, $\underline{\mathbf{V}}(k)$ is the $N_T \times N_T$ unitary matrix of right eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, and $\underline{\Lambda}(k)$ is the $N_R \times N_T$ diagonal matrix of singular values of $\hat{\underline{\mathbf{H}}}(k)$.

[1054] In another embodiment for performing channel eigen-decomposition, unit 224 first obtains a square matrix $\underline{\mathbf{R}}(k)$ as $\underline{\mathbf{R}}(k) = \hat{\underline{\mathbf{H}}}^H(k)\hat{\underline{\mathbf{H}}}(k)$. The eigenvalues of the square matrix $\underline{\mathbf{R}}(k)$ would then be the squares of the singular values of the matrix $\hat{\underline{\mathbf{H}}}(k)$, and the eigen-vectors of $\underline{\mathbf{R}}(k)$ would be the right eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, or $\underline{\mathbf{V}}(k)$. The decomposition of $\underline{\mathbf{R}}(k)$ to obtain the eigenvalues and eigen-vectors is known in the art and not described herein. Similarly, another square matrix $\underline{\mathbf{R}}'(k)$ may be obtained as $\underline{\mathbf{R}}'(k) = \hat{\underline{\mathbf{H}}}(k)\hat{\underline{\mathbf{H}}}^H(k)$. The eigenvalues of this square matrix $\underline{\mathbf{R}}'(k)$ would also be the squares of the singular values of $\hat{\underline{\mathbf{H}}}(k)$, and the eigen-vectors of $\underline{\mathbf{R}}'(k)$ would be the left eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, or $\underline{\mathbf{U}}(k)$.

[1055] The channel eigen-decomposition is used to decompose the MIMO channel into its eigenmodes, at frequency f_k, for each value of k where $k = (0, 1, \dots, N_F - 1)$. The rank $r(k)$ of $\hat{\underline{\mathbf{H}}}(k)$ corresponds to the number of eigenmodes for the MIMO channel at frequency f_k, which corresponds to the number of independent channels (i.e., the number of spatial subchannels) available in frequency bin k.

[1056] As described in further detail below, the columns of $\underline{\mathbf{V}}(k)$ are the steering vectors associated with frequency f_k to be used at the transmitter for the elements of the modulation symbol vectors, $\underline{\mathbf{s}}(n)$. Correspondingly, the columns of $\underline{\mathbf{U}}(k)$ are the steering vectors associated with frequency f_k to be used at the receiver for the elements of the received symbol vectors, $\underline{\mathbf{r}}(n)$. The matrices $\underline{\mathbf{U}}(k)$ and $\underline{\mathbf{V}}(k)$, for $k = (0, 1, \dots, N_F - 1)$, are used to orthogonalize the symbol streams transmitted on the

eigenmodes at each frequency f_k. When these matrices are used to precondition the modulation symbol streams at the transmitter and to condition the received symbol streams at the receiver, either in the frequency domain or the time domain, the result is the overall orthogonalization of the symbol streams. This then allows for separate coding/modulation per eigenmode (as opposed to per bin), which can greatly simplify the processing at the transmitter and receiver.

[1057] The elements along the diagonal of $\underline{\Lambda}(k)$ are $\lambda_{ii}(k)$, for $i = \{1, 2, \ldots, r(k)\}$, where $r(k)$ is the rank of $\underline{\hat{\mathbf{H}}}(k)$. The columns of $\underline{\mathbf{U}}(k)$ and $\underline{\mathbf{V}}(k)$, $\underline{\mathbf{u}}_i(k)$ and $\underline{\mathbf{v}}_i(k)$, respectively, are solutions to the eigen equation, which may be expressed as:

$$\underline{\hat{\mathbf{H}}}(k)\underline{\mathbf{v}}_i(k) = \lambda_{ii}\underline{\mathbf{u}}_i(k) \quad . \qquad \text{Eq (2)}$$

[1058] The three sets of matrices, $\underline{\mathbf{U}}(k)$, $\underline{\Lambda}(k)$, and $\underline{\mathbf{V}}(k)$, for $k = (0, 1, \ldots, N_F - 1)$, may be provided in two forms - a "sorted" form and a "random-ordered" form. In the sorted form, the diagonal elements of each matrix $\underline{\Lambda}(k)$ are sorted in decreasing order so that $\lambda_{11}(k) \geq \lambda_{22}(k) \geq \ldots \geq \lambda_{rr}(k)$, and their eigen-vectors are arranged in corresponding order in $\underline{\mathbf{U}}(k)$ and $\underline{\mathbf{V}}(k)$. The sorted form is indicated by the subscript *s*, i.e., $\underline{\mathbf{U}}_s(k)$, $\underline{\Lambda}_s(k)$, and $\underline{\mathbf{V}}_s(k)$, for $k = (0, 1, \ldots, N_F - 1)$.

[1059] In the random-ordered form, the ordering of the singular values and eigen-vectors may be random and further independent of frequency. The random form is indicated by the subscript *r*. The particular form selected for use, sorted or random-ordered, influences the selection of the eigenmodes for use for data transmission and the coding and modulation scheme to be used for each selected eigenmode.

[1060] A weight computation unit 230 receives the set of diagonal matrices, $\underline{\underline{\Lambda}}$, which contains a set of singular values (i.e., $\lambda_{11}(k), \lambda_{22}(k), \ldots, \lambda_{rr}(k)$) for each frequency bin. Weight computation unit 230 then derives a set of weighting matrices, $\underline{\underline{\mathbf{W}}}$, where $\underline{\underline{\mathbf{W}}} = [\underline{\mathbf{W}}(0) \ \ldots \ \underline{\mathbf{W}}(k) \ \ldots \ \underline{\mathbf{W}}(N_F - 1)]$. The weighting matrices are used to scale the modulation symbol vectors, $\underline{\mathbf{s}}(n)$, in either the time or frequency domain, as described below.

[1061] Weight computation unit 230 includes a channel inversion unit 232 and (optionally) a water-pouring analysis unit 234. Channel inversion unit 232 derives a set of weights, $\underline{\mathbf{w}}_{ii}$, for each eigenmode, which is used to combat the frequency selective fading on the eigenmode. Water-pouring analysis unit 234 derives a set of scaling values, $\underline{\mathbf{b}}$, for the eigenmodes of the MIMO channel. These scaling values are indicative of the transmit powers allocated to the eigenmodes. Channel inversion and water-pouring are described in further detail below.

Channel Inversion

[1062] **FIG. 3A** is a diagram that graphically illustrates the derivation of the set of weights, $\underline{\mathbf{w}}_{ii}$, used to invert the frequency response of each eigenmode. The set of diagonal matrices, $\underline{\Lambda}(k)$ for $k = (0, 1, \ldots, N_F - 1)$, is shown arranged in order along an axis 310 that represents the frequency dimension. The singular values, $\lambda_{ii}(k)$ for $i = (1, 2, \ldots, N_S)$, of each matrix $\underline{\Lambda}(k)$ are located along the diagonal of the matrix. Axis 312 may thus be viewed as representing the spatial dimension. Each eigenmode of the MIMO channel is associated with a set of elements, $\{\lambda_{ii}(k)\}$ for $k = (0, 1, \ldots, N_F - 1)$, that is indicative of the frequency response of that eigenmode. The set of elements $\{\lambda_{ii}(k)\}$ for each eigenmode is shown by the shaded boxes along a dashed line 314. For each eigenmode that experiences frequency selective fading, the elements $\{\lambda_{ii}(k)\}$ for the eigenmode may be different for different values of k.

[1063] Since frequency selective fading causes ISI, the deleterious effects of ISI may be mitigated by "inverting" each eigenmode such that it appears flat in frequency at the receiver. The channel inversion may be achieved by deriving a set of weights, $\{w_{ii}(k)\}$ for $k = (0, 1, \ldots, N_F - 1)$, for each eigenmode such that the product of the weights and the corresponding eigenvalues (i.e., the squares of the diagonal elements) are approximately constant for all values of k, which may be expressed as $w_{ii}(k) \cdot \lambda_{ii}^2(k) = a_i$, for $k = (0, 1, \ldots, N_F - 1)$.

[1064] For eigenmode i, the set of weights for the N_F frequency bins, $\underline{\mathbf{w}}_{ii} = [w_{ii}(0) \ \ldots \ w_{ii}(k) \ \ldots \ w_{ii}(N_F - 1)]^T$, used to invert the channel may be derived as:

$$w_{ii}(k) = \frac{a_i}{\lambda_{ii}^2(k)} \quad , \qquad \text{for } k = (0,\ 1,\ \ldots\ , N_F - 1)\ , \qquad \text{Eq (3)}$$

where a_i is a normalization factor that may be expressed as:

$$a_i = \frac{\sum_{k=0}^{N_F-1} \lambda_{ii}^2(k)}{\sum_{k=0}^{N_F-1} \frac{1}{\lambda_{ii}^2(k)}} \ . \qquad \text{Eq (4)}$$

As shown in equation (4), a normalization factor a_i is determined for each eigenmode based on the set of eigenvalues (i.e., the squared singular values), $\{\lambda_{ii}^2(k)\}$ for $k = (0,\ 1,\ \ldots\ , N_F - 1)$, associated with that eigenmode. The normalization factor a_i is defined such that $\sum_{k=0}^{N_F-1} w_{ii}(k) = \sum_{k=0}^{N_F-1} \lambda_{ii}^2(k)$.

[1065] **FIG. 3B** is a diagram that graphically illustrates the relationship between the set of weights for a given eigenmode and the set of eigenvalues for that eigenmode. For eigenmode *i*, the weight $w_{ii}(k)$ for each frequency bin is inversely related to the eigenvalue $\lambda_{ii}^2(k)$ for that bin, as shown in equation (3). To flatten the spatial subchannel and minimize or reduce ISI, it is undesirable to selectively eliminate transmit power on any frequency bin. The set of N_F weights for each eigenmode is used to scale the modulation symbols, $\underline{s}(n)$, in the frequency or time domain, prior to transmission on the eigenmode.

[1066] For the sorted order form, the singular values $\lambda_{ii}(k)$, for $i = (1,\ 2,\ \ldots\ , N_S)$, for each matrix $\underline{\Lambda}(k)$ are sorted such that the diagonal elements of $\underline{\Lambda}(k)$ with smaller indices are generally larger. Eigenmode 0 (which is often referred to as the principle eigenmode) would then be associated with the largest singular value in each of the N_F diagonal matrices, $\underline{\Lambda}(k)$, eigenmode 1 would then be associated with the second largest singular value in each of the N_F diagonal matrices, and so on. Thus, even though the channel inversion is performed over all N_F frequency bins for each eigenmode, the eigenmodes with lower indices are not likely to have too many bad bins (if any). Thus,

at least for eigenmodes with lower indices, excessive transmit power is not used for bad bins.

[1067] The channel inversion may be performed in various manners to invert the MIMO channel, and this is within the scope of the invention. In one embodiment, the channel inversion is performed for each eigenmode selected for use. In another embodiment, the channel inversion may be performed for some eigenmodes but not others. For example, the channel inversion may be performed for each eigenmode determined to induce excessive ISI. The channel inversion may also be dynamically performed for some or all eigenmodes selected for use, for example, when the MIMO channel is determined to be frequency selective (e.g., based on some defined criteria).

[1068] Channel inversion is described in further detail in U.S. Patent Application Serial No. 09/860,274, filed May 17, 2001, U.S. Patent Application Serial No. 09/881,610, filed June 14, 2001, and U.S. Patent Application Serial No. 09/892,379, filed June 26, 2001, all three entitled "Method and Apparatus for Processing Data for Transmission in a Multi-Channel Communication System Using Selective Channel Inversion," assigned to the assignee of the present application and incorporated herein by reference.

Water-Pouring

[1069] In an embodiment, water-pouring analysis is performed (if at all) across the spatial dimension such that more transmit power is allocated to eigenmodes with better transmission capabilities. The water-pouring power allocation is analogous to pouring a fixed amount of water into a vessel with an irregular bottom, where each eigenmode corresponds to a point on the bottom of the vessel, and the elevation of the bottom at any given point corresponds to the inverse of the SNR associated with that eigenmode. A low elevation thus corresponds to a high SNR and, conversely, a high elevation corresponds to a low SNR. The total available transmit power, P_{total}, is then "poured" into the vessel such that the lower points in the vessel (i.e., those with higher SNRs) are filled first, and the higher points (i.e., those with lower SNRs) are filled later. A constant P_{set} is indicative of the water surface level for the vessel after all of the total available transmit power has been poured. This constant may be estimated initially based on various system parameters. The power allocation is dependent on the total available transmit power and the depth of the vessel over the bottom surface. The

points with elevations above the water surface level are not filled (i.e., eigenmodes with SNRs below a particular value are not used for data transmission).

[1070] In an embodiment, the water-pouring is not performed across the frequency dimension because this tends to exaggerate the frequency selectivity of the eigenmodes created by the channel eigenmode decomposition described above. The water-pouring may be performed such that all eigenmodes are used for data transmission, or only a subset of the eigenmodes is used (with bad eigenmodes being discarded). It can be shown that water-pouring across the eigenmodes, when used in conjunction with the channel inversion with the singular values sorted in descending order, can provide near-optimum performance while mitigating the need for equalization at the receiver.

[1071] The water-pouring may be performed by water-pouring analysis unit 234 as follows. Initially, the total power in each eigenmode is determined as:

$$P_{i,\lambda} = \sum_{k=0}^{N_F-1} \lambda_{ii}^2(k) \quad . \qquad \text{Eq (5)}$$

[1072] The SNR for each eigenmode may then be determined as:

$$\text{SNR}_i = \frac{P_{i,\lambda}}{\sigma^2} \quad , \qquad \text{Eq (6)}$$

where σ^2 is the received noise variance, which may also be denoted as the received noise power N_0. The received noise power corresponds to the noise power on the recovered symbols at the receiver, and is a parameter that may be provided by the receiver to the transmitter as part of the reported CSI.

[1073] The transmit power, P_i, to be allocated to each eigenmode may then be determined as:

$$P_i = \max\left[\left(P_{set} - \frac{1}{\text{SNR}_i}\right), 0\right] \quad , \text{ and} \qquad \text{Eq (7a)}$$

$$P_{total} \geq \sum_{i=1}^{N_S} P_i \quad , \qquad \text{Eq (7b)}$$

where P_{set} is a constant that may be derived from various system parameters, and P_{total} is the total transmit power available for allocation to the eigenmodes.

[1074] As shown in equation (7a), each eigenmode of sufficient quality is allocated transmit power of $\left(P_{set} - \frac{1}{\text{SNR}_i}\right)$. Thus, eigenmodes that achieve better SNRs are allocated more transmit powers. The constant P_{set} determines the amounts of transmit power to allocate to the better eigenmodes. This then indirectly determines which eigenmodes get selected for use since the total available transmit power is limited and the power allocation is constrained by equation (7b).

[1075] Water-pouring analysis unit 234 thus receives the set of diagonal matrices, $\underline{\underline{\Lambda}}$, and the received noise power, σ^2. The matrices $\underline{\underline{\Lambda}}$ are then used in conjunction with the received noise power to derive a vector of scaling values, $\underline{\mathbf{b}} = [b_0 \ \ldots \ b_i \ \ldots \ b_{N_S}]^T$, where $b_i = P_i$ for $i = (1, 2, \ldots, N_S)$. The P_i are the solutions to the water-pouring equations (7a) and (7b). The scaling values in $\underline{\mathbf{b}}$ are indicative of the transmit powers allocated to the N_S eigenmodes, where zero or more eigenmodes may be allocated no transmit power.

[1076] **FIG. 4** is a flow diagram of an embodiment of a process 400 for allocating the total available transmit power to a set of eigenmodes. Process 400, which is one specific water-pouring implementation, determines the transmit powers, P_i, for $i \in I$, to be allocated to the eigenmodes in set I, given the total transmit power, P_{total}, available at the transmitter, the set of eigenmode total powers, $P_{i,\lambda}$, and the received noise power, σ^2.

[1077] Initially, the variable n used to denote the iteration number is set to one (i.e., $n = 1$) (step 412). For the first iteration, set $I(n)$ is defined to include all of the eigenmodes for the MIMO channel, or $I(n) = \{1, 2, \ldots, N_S\}$ (step 414). The cardinality (or length) of set $I(n)$ for the current iteration n is then determined as $L_I(n) = |I(n)|$, which is $L_I(n) = N_S$ for the first iteration (step 416).

[1078] The total effective power, $P_{eff}(n)$, to be distributed across the eigenmodes in set $I(n)$ is next determined (step 418). The total effective power is defined to be equal

to the total available transmit power, P_{total}, plus the sum of the inverse SNRs for the eigenmodes in set $I(n)$. This may be expressed as:

$$P_{eff}(n) = P_{total} + \sum_{i \in I(n)} \frac{\sigma^2}{P_{i,\lambda}} \quad . \qquad \text{Eq (8)}$$

[1079] The total available transmit power is then allocated to the eigenmodes in set $I(n)$. The index i used to iterate through the eigenmodes in set $I(n)$ is initialized to one (i.e., $i = 1$) (step 420). The amount of transmit power to allocate to eigenmode i is then determined (step 422) based on the following:

$$P_i(n) = \frac{P_{eff}(n)}{L_I(n)} - \frac{\sigma^2}{P_{i,\lambda}} \quad . \qquad \text{Eq (9)}$$

Each eigenmode in set $I(n)$ is allocated transmit power, P_i, in step 422. Steps 424 and 426 are part of a loop to allocate transmit power to each of the eigenmodes in set $I(n)$.

[1080] **FIG. 5A** graphically illustrates the total effective power, P_{eff}, for an example MIMO system with three eigenmodes. Each eigenmode has an inverse SNR equal to $\sigma^2 / \lambda_{ii}^2$, for $i = \{1, 2, 3\}$, which assumes a normalized transmit power of 1.0. The total transmit power available at the transmitter is $P_{total} = P_1 + P_2 + P_3$, and is represented by the shaded area in FIG. 5A. The total effective power is represented by the area in the shaded and unshaded regions in FIG. 5A.

[1081] For water-pouring, although the bottom of the vessel has an irregular surface, the water level at the top remains constant across the vessel. Likewise and as shown in FIG. 5A, after the total available transmit power, P_{total}, has been distributed to the eigenmodes, the final power level is constant across all eigenmodes. This final power level is determined by dividing $P_{eff}(n)$ by the number of eigenmodes in set $I(n)$, $L_I(n)$. The amount of power allocated to eigenmode i is then determined by subtracting the inverse SNR of that eigenmode, $\sigma^2 / \lambda_{ii}^2$, from the final power level, $P_{eff}(n) / L_I(n)$, as given by equation (9) and shown in FIG. 5A.

[1082] **FIG. 5B** shows a situation whereby the water-pouring power allocation results in an eigenmode receiving negative power. This occurs when the inverse SNR

of the eigenmode is above the final power level, which is expressed by the condition $(P_{eff}(n)/L_I(n)) < (\sigma^2/\lambda_{ii}^2)$.

[1083] Referring back to FIG. 4, at the end of the power allocation, a determination is made whether or not any eigenmode has been allocated negative power (i.e., $P_i < 0$) (step 428). If the answer is yes, then the process continues by removing from set $I(n)$ all eigenmodes that have been allocated negative powers (step 430). The index n is incremented by one (i.e., $n = n+1$) (step 432). The process then returns to step 416 to allocate the total available transmit power among the remaining eigenmodes in set $I(n)$. The process continues until all eigenmodes in set $I(n)$ have been allocated positive transmit powers, as determined in step 428. The eigenmodes not in set $I(n)$ are allocated zero power.

[1084] Water-pouring is also described by Robert G. Gallager, in "Information Theory and Reliable Communication," John Wiley and Sons, 1968, which is incorporated herein by reference. A specific algorithm for performing the basic water-pouring process for a MIMO-OFDM system is described in U.S. Patent Application Serial No. 09/978,337, entitled "Method and Apparatus for Determining Power Allocation in a MIMO Communication System," filed October 15, 2001. Water-pouring is also described in further detail in U.S. Patent Application Serial No. 10/056,275, entitled "Reallocation of Excess Power for Full Channel-State Information (CSI) Multiple-Input, Multiple-Output (MIMO) Systems," filed January 23, 2002. These applications are assigned to the assignee of the present application and incorporated herein by reference.

[1085] If water-pouring is performed to allocate the total available transmit power to the eigenmodes, then water-pouring analysis unit 234 provides a set of N_S scaling values, $\underline{\mathbf{b}} = \{b_0\ \ldots\ b_i\ \ldots\ b_{N_S}\}$, for the N_S eigenmodes. Each scaling value is for a respective eigenmode and is used to scale the set of weights determined for that eigenmode.

[1086] For eigenmode i, a set of weights, $\underline{\hat{\mathbf{w}}}_{ii} = [\hat{w}_{ii}(0)\ \ldots\ \hat{w}_{ii}(k)\ \ldots\ \hat{w}_{ii}(N_F-1)]^T$, used to invert the channel and scale the transmit power of the eigenmode may be derived as:

$$\hat{w}_{ii}(k) = \frac{a_i b_i}{\lambda_{ii}^2(k)} \, , \qquad \text{for } k = (0,\ 1,\ \ldots\ , N_F - 1) \, , \qquad \text{Eq (10)}$$

where the normalization factor, a_i, and the scaling value, b_i, are derived as described above.

[1087] Weight computation unit 230 provides the set of weighting matrices, $\underline{\underline{\mathbf{W}}}$, which may be obtained using the weights $w_{ii}(k)$ or $\hat{w}_{ii}(k)$. Each weighting matrix, $\underline{\mathbf{W}}(k)$, is a diagonal matrix whose diagonal elements are composed of the weights derived above. In particular, if only channel inversion is performed, then each weighting matrix, $\underline{\mathbf{W}}(k)$, for $k = (0,\ 1,\ \ldots\ , N_F - 1)$, is defined as:

$$\underline{\mathbf{W}}(k) = \operatorname{diag}\left(w_{11}(k),\ w_{22}(k),\ \ldots\ , w_{N_S N_S}(k)\right) \, , \qquad \text{Eq (11a)}$$

where $w_{ii}(k)$ is derived as shown in equation (3). And if both channel inversion and water-pouring are performed, then each weighting matrix, $\underline{\mathbf{W}}(k)$, for $k = (0,\ 1,\ \ldots\ , N_F - 1)$, is defined as:

$$\underline{\mathbf{W}}(k) = \operatorname{diag}\left(\hat{w}_{11}(k),\ \hat{w}_{22}(k),\ \ldots\ , \hat{w}_{N_S N_S}(k)\right) \, , \qquad \text{Eq (11b)}$$

where $\hat{w}_{ii}(k)$ is derived as shown in equation (10).

[1088] Referring back to FIG. 2, a scaler/IFFT 236 receives (1) the set of unitary matrices, $\underline{\underline{\mathbf{V}}}$, which are the matrices of right eigen-vectors of $\underline{\underline{\hat{\mathbf{H}}}}$, and (2) the set of weighting matrices, $\underline{\underline{\mathbf{W}}}$, for all N_F frequency bins. Scaler/IFFT 236 then derives a spatio-temporal pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, for the transmitter based on the received matrices. Initially, the square root of each weighting matrix, $\underline{\mathbf{W}}(k)$, is computed to obtain a corresponding matrix, $\sqrt{\underline{\mathbf{W}}(k)}$, whose elements are the square roots of the elements of $\underline{\mathbf{W}}(k)$. The elements of the weighting matrices, $\underline{\mathbf{W}}(k)$ for $k = (0,\ 1,\ \ldots\ , N_F - 1)$, are related to the power of the eigenmodes. The square root then transforms the power to the equivalent signal scaling. For each frequency bin k, the product of the square-root weighting matrix, $\sqrt{\underline{\mathbf{W}}(k)}$, and the corresponding unitary

matrix, $\underline{\mathbf{V}}(k)$, is then computed to provide a product matrix, $\underline{\mathbf{V}}(k)\sqrt{\underline{\mathbf{W}}(k)}$. The set of product matrices, $\underline{\mathbf{V}}(k)\sqrt{\underline{\mathbf{W}}(k)}$ for $k = (0, 1, \ldots, N_F - 1)$, which is also denoted as $\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}$, defines the optimal or near-optimal spatio-spectral shaping to be applied to the modulation symbol vectors, $\underline{\mathbf{s}}(n)$.

[1089] An inverse FFT of $\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}$ is then computed to derive the spatio-temporal pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, for the transmitter, which may be expressed as:

$$\underline{\mathbf{P}}_{tx}(n) = \mathrm{IFFT}\,[\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}] \quad . \qquad \text{Eq (12)}$$

The pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, is an $N_T \times N_T$ matrix. Each element of $\underline{\mathbf{P}}_{tx}(n)$ is a set of N_F time-domain values, which is obtained by the inverse FFT of a set of values for the corresponding element of the product matrices, $\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}$. Each column of $\underline{\mathbf{P}}_{tx}(n)$ is a steering vector for a corresponding element of $\underline{\mathbf{s}}(n)$.

[1090] A convolver 240 receives and preconditions the modulation symbol vectors, $\underline{\mathbf{s}}(n)$, with the pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, to provide the transmitted symbol vectors, $\underline{\mathbf{x}}(n)$. In the time domain, the preconditioning is a convolution operation, and the convolution of $\underline{\mathbf{s}}(n)$ with $\underline{\mathbf{P}}_{tx}(n)$ may be expressed as:

$$\underline{\mathbf{x}}(n) = \sum_{\ell} \underline{\mathbf{P}}_{tx}(\ell)\underline{\mathbf{s}}(n-\ell) \quad . \qquad \text{Eq (13)}$$

The matrix convolution shown in equation (13) may be performed as follows. To derive the i-th element of the vector $\underline{\mathbf{x}}(n)$ for time n, $x_i(n)$, the inner product of the i-th row of the matrix $\underline{\mathbf{P}}_{tx}(\ell)$ with the vector $\underline{\mathbf{s}}(n-\ell)$ is formed for a number of delay indices (e.g., $0 \le \ell \le (N_F - 1)$), and the results are accumulated to derive the element $x_i(n)$. The preconditioned symbol streams transmitted on each transmit antenna (i.e., each element of $\underline{\mathbf{x}}(n)$ or $x_i(n)$) is thus formed as a weighted combination of the N_R modulation symbol streams, with the weighting determined by the appropriate column of the matrix $\underline{\mathbf{P}}_{tx}(n)$. The process is repeated such that each element of $\underline{\mathbf{x}}(n)$ is obtained from a respective column of the matrix $\underline{\mathbf{P}}_{tx}(n)$ and the vector $\underline{\mathbf{s}}(n)$.

[1091] Each element of $\underline{\mathbf{x}}(n)$ corresponds to a sequence of preconditioned symbols to be transmitted over a respective transmit antenna. The N_T preconditioned symbol sequences collectively form a sequence of vectors, which are also referred to as the transmitted symbol vectors, $\underline{\mathbf{x}}(n)$, with each such vector including up to N_T preconditioned symbols to be transmitted from up to N_T transmit antennas for the n-th symbol period. The N_T preconditioned symbol sequences are provided to transmitters 122a through 122t and processed to derive N_T modulated signals, which are then transmitted from antennas 124a through 124t, respectively.

[1092] The embodiment shown in FIG. 2 performs time-domain beam-steering of the modulation symbol vectors, $\underline{\mathbf{s}}(n)$. The beam-steering may also be performed in the frequency domain. This can be done using techniques, such as the overlap-add method, which are well-known in the digital signal processing field, for implementing finite-duration impulse response (FIR) filters in the frequency domain. In this case, the sequences that make up the elements of the matrix $\underline{\mathbf{P}}_{tx}(n)$ for $n = (0,\ 1,\ \ldots\ , N_F - 1)$ are each padded with $N_O - N_F$ zeros, resulting in a matrix of zero-padded sequences, $\underline{\mathbf{q}}_{tx}(n)$, for $n = (0,\ 1,\ \ldots\ , N_O - 1)$. An N_O-point fast Fourier transform (FFT) is then computed for each zero-padded sequence in the matrix $\underline{\mathbf{q}}_{tx}(n)$, resulting in a matrix $\underline{\mathbf{Q}}_{tx}(k)$ for $k = (0,\ 1,\ \ldots\ , N_O - 1)$.

[1093] Also, the sequences of modulation symbols that make up the elements of $\underline{\mathbf{s}}(n)$ are each broken up into subsequences of length $N_{SS} = N_O - N_F + 1$. Each subsequence is then padded with $N_F - 1$ zeros to provide a corresponding vector of length N_O. The sequences of $\underline{\mathbf{s}}(n)$ are thus processed to provide sequences of vectors of length N_O, $\tilde{\underline{\mathbf{s}}}_\ell(n)$, where the subscript ℓ is the index for the vectors that correspond to the zero-padded subsequences. An N_O-point fast Fourier transform is then computed for each of the zero-padded subsequences, resulting in a sequence of frequency-domain vectors, $\tilde{\underline{\mathbf{S}}}_\ell(k)$, for different values of ℓ. Each vector $\tilde{\underline{\mathbf{S}}}_\ell(k)$, for a given ℓ, includes a set of frequency-domain vectors of length N_O (i.e., for $k = (0,\ 1,\ \ldots\ , N_O - 1)$). The matrix $\underline{\mathbf{Q}}_{tx}(k)$ is then multiplied with the vector $\tilde{\underline{\mathbf{S}}}_\ell(k)$, for each value of ℓ, where the pre-multiplication is performed for each value of k, i.e., for $k = (0,\ 1,\ \ldots\ , N_O - 1)$.

The inverse FFTs are then computed for the matrix-vector product $\underline{\underline{\mathbf{Q}}}_{tx}(k)\underline{\tilde{\mathbf{S}}}_{\ell}(k)$ to provide a set of time-domain subsequences of length N_O. The resulting subsequences are then reassembled, according to the overlap-add method, or similar means, as is well-known in the art, to form the desired output sequences.

[1094] **FIG. 6** is a flow diagram of an embodiment of a process 600 that may be performed at the transmitter unit to implement the various transmit processing techniques described herein. Initially, data to be transmitted (i.e., the information bits) is processed in accordance with a particular processing scheme to provide a number of streams of modulation symbols (step 612). As noted above, the processing scheme may include one or more coding schemes and one or more modulation schemes (e.g., a separate coding and modulation scheme for each modulation symbol stream).

[1095] An estimated channel response matrix for the MIMO channel is then obtained (step 614). This matrix may be the estimated channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$, or the estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$, which may be provided to the transmitter from the receiver. The estimated channel response matrix is then decomposed (e.g., using channel eigen-decomposition) to obtain a set of matrices of right eigen-vectors, $\underline{\underline{\mathbf{V}}}$, and a set of matrices of singular values, $\underline{\underline{\Lambda}}$ (step 616).

[1096] A number of sets of weights, $\underline{\mathbf{w}}_{ii}$, are then derived based on the matrices of singular values (step 618). One set of weight may be derived for each eigenmode used for data transmission. These weights are used to reduce or minimize intersymbol interference at the receiver by inverting the frequency response of each eigenmode selected for use.

[1097] A set of scaling values, $\underline{\mathbf{b}}$, may also be derived based on the matrices of singular values (step 620). Step 620 is optional, as indicated by the dashed box for step 620 in FIG. 6. The scaling values may be derived using water-pouring analysis and are used to adjust the transmit powers of the selected eigenmodes.

[1098] A pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, is then derived based on the matrices of right eigen-vectors, $\underline{\underline{\mathbf{V}}}$, the sets of weights, $\underline{\mathbf{w}}_{ii}$, and (if available) the set of scaling values, $\underline{\mathbf{b}}$ (step 622). The streams of modulation symbols are then preconditioned (in either the time domain or frequency domain) based on the pulse-shaping matrix to

provide a number of streams of preconditioned symbols, $\underline{\mathbf{x}}(n)$, to be transmitted over the MIMO channel (step 624).

[1099] Time-domain transmit processing with channel eigenmode decomposition and water-pouring is described in further detail in U.S. Patent Application Serial No. 10/017,038, entitled "Time-Domain Transmit and Receive Processing with Channel Eigen-mode Decomposition for MIMO Systems," filed December 7, 2001, which is assigned to the assignee of the present application and incorporated herein by reference.

[1100] **FIG. 7** is a block diagram of an embodiment of a receiver unit 700 capable of implementing various processing techniques described herein. Receiver unit 700 is an embodiment of the receiver portion of receiver system 150 in FIG. 1. Receiver unit 700 includes (1) a RX MIMO processor 160a that processes N_R received symbol streams to provide N_T recovered symbol streams and (2) a RX data processor 162a that demodulates, deinterleaves, and decodes the recovered symbols to provide decoded bits. RX MIMO processor 160a and RX data processor 162a are one embodiment of RX MIMO processor 160 and RX data processor 162, respectively, in FIG. 1.

[1101] Referring back to FIG. 1, the transmitted signals from N_T transmit antennas are received by each of N_R antennas 152a through 152r. The received signal from each antenna is routed to a respective receiver 154, which is also referred to as a front-end processor. Each receiver 154 conditions (e.g., filters, amplifies, and frequency downconverts) a respective received signal, and further digitizes the conditioned signal to provide ADC samples. Each receiver 154 may further data demodulate the ADC samples with a recovered pilot to provide a respective stream of received symbols. Receivers 154a through 154r thus provide N_R received symbol streams. These streams collectively form a sequence of vectors, which are also referred to as the received symbol vectors, $\underline{\mathbf{r}}(n)$, with each such vector including N_R received symbols from the N_R receivers 154 for the n-th symbol period. The received symbol vectors, $\underline{\mathbf{r}}(n)$, are then provided to RX MIMO processor 160a.

[1102] Within RX MIMO processor 160a, a channel estimator 712 receives the vectors $\underline{\mathbf{r}}(n)$ and derives an estimated channel impulse response matrix, $\hat{\underline{\underline{\mathcal{H}}}}$, which may be sent back to the transmitter system and used in the transmit processing. An FFT 714

then performs an FFT on the estimated channel impulse response matrix, $\hat{\underline{\underline{\mathcal{H}}}}$, to obtain the estimated channel frequency response matrix, $\hat{\underline{\underline{\mathbf{H}}}}$ (i.e., $\hat{\underline{\underline{\mathbf{H}}}} = \text{FFT}[\hat{\underline{\underline{\mathcal{H}}}}]$).

[1103] A unit 716 then performs the channel eigen-decomposition of $\hat{\underline{\mathbf{H}}}(k)$, for each frequency bin k, to obtain the corresponding matrix of left eigen-vectors, $\underline{\mathbf{U}}(k)$. Each column of $\underline{\underline{\mathbf{U}}}$, where $\underline{\underline{\mathbf{U}}} = [\underline{\mathbf{U}}(0) \;...\; \underline{\mathbf{U}}(k) \;...\; \underline{\mathbf{U}}(N_F - 1)]$, is a steering vector for a corresponding element of $\underline{\mathbf{r}}(n)$, and is used to orthogonalize the received symbol streams. An IFFT 718 then performs the inverse FFT of $\underline{\underline{\mathbf{U}}}$ to obtain a spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}(n)$, for the receiver system.

[1104] A convolver 720 then conditions the received symbol vectors, $\underline{\mathbf{r}}(n)$, with the conjugate transpose of the spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}^H(n)$, to obtain the recovered symbol vectors, $\hat{\underline{\mathbf{s}}}(n)$, which are estimates of the modulation symbol vectors, $\underline{\mathbf{s}}(n)$. In the time domain, the conditioning is a convolution operation, which may be expressed as:

$$\hat{\underline{\mathbf{s}}}(n) = \sum_{\ell} \underline{\mathcal{U}}^H(\ell)\, \underline{\mathbf{r}}(n-\ell) \quad . \qquad \text{Eq (14)}$$

[1105] The pulse-shaping at the receiver may also be performed in the frequency domain, similar to that described above for the transmitter. In this case, the N_R sequences of received symbols for the N_R receive antennas, which make up the sequence of received symbol vectors, $\underline{\mathbf{r}}(n)$, are each broken up into subsequences of N_{SS} received symbols, and each subsequence is zero-padded to provide a corresponding vector of length N_O. The N_R sequences of $\underline{\mathbf{r}}(n)$ are thus processed to provide N_R sequences of vectors of length N_O, $\tilde{\underline{\mathbf{r}}}_\ell(n)$, where the subscript ℓ is the index for the vectors that correspond to the zero-padded subsequences. Each zero-padded subsequence is then transformed via an FFT, resulting in a sequence of frequency-domain vectors, $\underline{\mathbf{R}}_\ell(k)$, for different values of ℓ. Each vector $\underline{\mathbf{R}}_\ell(k)$, for a given ℓ, includes a set of frequency-domain vectors of length N_O (i.e., for $k = (0,\ 1,\ \ldots\ , N_O - 1)$).

[1106] The conjugate transpose of the spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}^H(n)$, is also zero-padded and transformed via an FFT to obtain a frequency-domain

matrix, $\underline{\tilde{\mathbf{U}}}^H(k)$ for $k=(0,\ 1,\ \ldots\ ,N_O-1)$. The vector $\underline{\mathbf{R}}_\ell(k)$, for each value of ℓ, is then pre-multiplied with the conjugate transpose matrix $\underline{\tilde{\mathbf{U}}}^H(k)$ (where the pre-multiplication is performed for each value of k, i.e., for $k=(0,\ 1,\ \ldots\ ,N_O-1)$) to obtain a corresponding frequency-domain vector $\underline{\hat{\mathbf{S}}}_\ell(k)$. Each vector $\underline{\hat{\mathbf{S}}}_\ell(k)$, which includes a set of frequency-domain vectors of length N_O, can then be transformed via an inverse FFT to provide a corresponding set of time-domain subsequences of length N_O. The resulting subsequences are then reassembled according to the overlap-add method, or similar means, as is well-known in the art, to obtain sequences of recovered symbols, which corresponds to the set of recovered symbol vectors, $\underline{\hat{\mathbf{s}}}(n)$.

[1107] Thus recovered symbol vectors, $\underline{\hat{\mathbf{s}}}(n)$, may be characterized as a convolution in the time domain, as follows:

$$\underline{\hat{\mathbf{s}}}(n)=\sum_\ell \underline{\Gamma}(\ell)\,\underline{\mathbf{s}}(n-\ell)+\underline{\hat{\mathbf{z}}}(n)\ , \qquad \text{Eq (15)}$$

where $\underline{\Gamma}(\ell)$ is the inverse FFT of $\underline{\hat{\Lambda}}(k)=\underline{\Lambda}(k)\sqrt{\underline{\mathbf{W}}(k)}$; and

$\underline{\hat{\mathbf{z}}}(n)$ is the received noise as transformed by the receiver spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}^H(\ell)$.

The matrix $\underline{\Gamma}(n)$ is a diagonal matrix of eigen-pulses derived from the set of matrices $\underline{\underline{\hat{\Lambda}}}$, where $\underline{\underline{\hat{\Lambda}}}=[\underline{\hat{\Lambda}}(0)\ \ldots\ \underline{\hat{\Lambda}}(k)\ \ldots\ \underline{\hat{\Lambda}}(N_F-1)]$. In particular, each diagonal element of $\underline{\Gamma}(n)$ corresponds to an eigen-pulse that is obtained as the IFFT of a set of singular values, $[\hat{\lambda}_{ii}(0)\ \ldots\ \hat{\lambda}_{ii}(k)\ \ldots\ \hat{\lambda}_{ii}(N_F-1)]^T$, for a corresponding element of $\underline{\underline{\hat{\Lambda}}}$,

[1108] The two forms for ordering the singular values, sorted and random-ordered, result in two different types of eigen-pulses. For the sorted form, the resulting eigen-pulse matrix, $\underline{\Gamma}_s(n)$, is a diagonal matrix of pulses that are sorted in descending order of energy content. The pulse corresponding to the first diagonal element of the eigen-pulse matrix, $\{\underline{\Gamma}_s(n)\}_{11}$, has the most energy, and the pulses corresponding to elements further down the diagonal have successively less energy. Furthermore, when the SNR is low enough that water-pouring results in some of the frequency bins having little or no energy, the energy is taken out of the smallest eigen-pulses first. Thus, at low SNRs,

one or more of the eigen-pulses may have little or no energy. This has the advantage that at low SNRs, the coding and modulation are simplified through the reduction in the number of orthogonal subchannels. However, in order to approach channel capacity, the coding and modulation are performed separately for each eigen-pulse.

[1109] The random-ordered form of the singular values in the frequency domain may be used to further simplify coding and modulation (i.e., to avoid the complexity of separate coding and modulation for each element of the eigen-pulse matrix). In the random-ordered form, for each frequency bin, the ordering of the singular values is random instead of being based on their magnitude or size. This random ordering can result in approximately equal energy in all of the eigen-pulses. When the SNR is low enough to result in frequency bins with little or no energy, these bins are spread approximately evenly among the eigenmodes so that the number of eigen-pulses with non-zero energy is the same independent of SNR. At high SNRs, the random-order form has the advantage that all of the eigen-pulses have approximately equal energy, in which case separate coding and modulation for different eigenmodes are not required.

[1110] If the response of the MIMO channel is frequency selective, then the elements in the diagonal matrices, $\underline{\Lambda}(k)$, are time-dispersive. However, because of the pre-processing at the transmitter to invert the channel, the resulting recovered symbol sequences, $\underline{\hat{s}}(n)$, have little intersymbol interference, if the channel inversion is effectively performed. In that case, additional equalization would not be required at the receiver to achieve high performance.

[1111] If the channel inversion is not effective (e.g., due to an inaccurate estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$) then an equalizer may be used to equalize the recovered symbols, $\underline{\hat{s}}(n)$, prior to the demodulation and decoding. Various types of equalizer may be used to equalize the recovered symbol streams, including a minimum mean square error linear equalizer (MMSE-LE), a decision feedback equalizer (DFE), a maximum likelihood sequence estimator (MLSE), and so on.

[1112] Since the orthogonalization process at the transmitter and receiver results in decoupled (i.e., orthogonal) recovered symbol streams at the receiver, the complexity of the equalization required for the decoupled symbol streams is greatly reduced. In particular, the equalization may be achieved by parallel time-domain equalization of the independent symbol streams. The equalization may be performed as described in the

aforementioned U.S. Patent Application Serial No. 10/017,038, and in U.S. Patent Application Serial No. 09/993,087, entitled "Multiple-Access Multiple-Input Multiple-Output (MIMO) Communication System," filed November 6, 2001, which is assigned to the assignee of the present application and incorporated herein by reference.

[1113] For the embodiment in FIG. 7, the recovered symbol vectors, $\hat{\underline{s}}(n)$, are provided to RX data processor 162a. Within processor 162a, a symbol demapping element 732 demodulates each recovered symbol in $\hat{\underline{s}}(n)$ in accordance with a demodulation scheme that is complementary to the modulation scheme used for that symbol at the transmitter system. The demodulated data from symbol demapping element 732 is then de-interleaved by a deinterleaver 734. The deinterleaved data is further decoded by a decoder 736 to obtain the decoded bits, $\hat{d}_i$, which are estimates of the transmitted information bits, d_i. The deinterleaving and decoding are performed in a manner complementary to the interleaving and encoding, respectively, performed at the transmitter system. For example, a Turbo decoder or a Viterbi decoder may be used for decoder 736 if Turbo or convolutional coding, respectively, is performed at the transmitter system.

[1114] **FIG. 8** is a flow diagram of a process 800 that may be performed at the receiver unit to implement the various receive processing techniques described herein. Initially, an estimated channel response matrix for the MIMO channel is obtained (step 812). This matrix may be the estimated channel impulse response matrix, $\hat{\underline{\underline{\mathcal{H}}}}$, or the estimated channel frequency response matrix, $\hat{\underline{\underline{\mathbf{H}}}}$. The matrix $\hat{\underline{\underline{\mathcal{H}}}}$ or $\hat{\underline{\underline{\mathbf{H}}}}$ may be obtained, for example, based on pilot symbols transmitted over the MIMO channel. The estimated channel response matrix is then decomposed (e.g., using channel eigen-decomposition) to obtain a set of matrices of left eigen-vectors, $\underline{\underline{\mathbf{U}}}$ (step 814).

[1115] A pulse-shaping matrix $\underline{\mathcal{U}}(n)$ is then derived based on the matrices of left eigen-vectors, $\underline{\underline{\mathbf{U}}}$ (step 816). The streams of received symbols are then conditioned (in either the time domain or frequency domain) based on the pulse-shaping matrix $\underline{\mathcal{U}}(n)$ to provide the streams of recovered symbols (step 818). The recovered symbols are further processed in accordance with a particular receive processing scheme, which is

complementary to the transmit processing scheme used at the transmitter, to provide the decoded data (step 820).

[1116] Time-domain receive processing with channel eigenmode decomposition is described in further detail in the aforementioned U.S. Patent Application Serial No. 10/017,038.

[1117] The techniques for processing a data transmission at a transmitter and a receiver described herein may be implemented in various wireless communication systems, including but not limited to MIMO and CDMA systems. These techniques may also be used for the forward link and/or the reverse link.

[1118] The techniques described herein to process a data transmission at the transmitter and receiver may be implemented by various means. For example, these techniques may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the elements used to perform various signal processing steps at the transmitter (e.g., to code and modulate the data, decompose the channel response matrix, derive the weights to invert the channel, derive the scaling values for power allocation, derive the transmitter pulse-shaping matrix, precondition the modulation symbols, and so on) or at the receiver (e.g., to decompose the channel response matrix, derive the receiver pulse-shaping matrix, condition the received symbols, demodulate and decode the recovered symbols, and so on) may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[1119] For a software implementation, some or all of the signal processing steps at each of the transmitter and receiver may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memories 132 and 172 in FIG. 1) and executed by a processor (e.g., controllers 130 and 170). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as is known in the art.

[1120] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various

modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1121] **WHAT IS CLAIMED IS:**

CLAIMS

1. In a multiple-input multiple-output (MIMO) communication system, a method for processing data for transmission over a MIMO channel, comprising:

processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

deriving a pulse-shaping matrix based on an estimated response of the MIMO channel and in a manner to reduce intersymbol interference at a receiver; and

preconditioning the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

2. The method of claim 1, further comprising:

deriving a plurality of weights based on an estimated channel response matrix for the MIMO channel, wherein the weights are used to invert a frequency response of the MIMO channel, and wherein the pulse-shaping matrix is further derived based on the weights.

3. The method of claim 2, further comprising:

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors and a plurality of matrices of singular values, and

wherein the weights are derived based on the matrices of singular values and the pulse-shaping matrix is further derived based on the matrices of eigen-vectors.

4. The method of claim 2, wherein the estimated channel response matrix is descriptive of a plurality of eigenmodes of the MIMO channel.

5. The method of claim 4, wherein one set of weights is derived for each eigenmode used for data transmission and wherein the weights in each set are derived to invert the frequency response of the corresponding eigenmode.

6. The method of claim 4, further comprising:

deriving a plurality of scaling values based on the matrices of singular values, wherein the scaling values are used to adjust transmit powers for the eigenmodes of the MIMO channel, and wherein the pulse-shaping matrix is further derived based on the scaling values.

7. The method of claim 6, wherein the scaling values are derived based on water-pouring analysis.

8. The method of claim 3, wherein the estimated channel response matrix is provided in the frequency domain and is decomposed in the frequency domain.

9. The method of claim 3, wherein the estimated channel response matrix is decomposed using channel eigen-decomposition.

10. The method of claim 4, wherein eigenmodes associated with transmission capabilities below a particular threshold are not used for data transmission.

11. The method of claim 3, wherein the singular values in each matrix are sorted based on their magnitude.

12. The method of claim 4, wherein the singular values in each matrix are randomly ordered such that the eigenmodes of the MIMO channel are associated with approximately equal transmission capabilities.

13. The method of claim 1, wherein the pulse-shaping matrix comprises a plurality of sequences of time-domain values, and wherein the preconditioning is performed in the time domain by convolving the streams of modulation symbols with the pulse-shaping matrix.

14. The method of claim 1, wherein the pulse-shaping matrix comprises a plurality of sequences of frequency-domain values, and wherein the preconditioning is performed in the frequency domain by multiplying a plurality of streams of transformed modulation symbols with the pulse-shaping matrix.

15. The method of claim 1, wherein the pulse-shaping matrix is derived to maximize capacity by allocating more transmit power to eigenmodes of the MIMO channel having greater transmission capabilities.

16. The method of claim 1, wherein the pulse-shaping matrix is derived to provide approximately equal received signal-to-noise-and-interference ratios (SNRs) for the plurality of modulation symbol streams at the receiver.

17. The method of claim 1, wherein the particular processing scheme defines a separate coding and modulation scheme for each modulation symbol stream.

18. The method of claim 1, wherein the particular processing scheme defines a common coding and modulation scheme for all modulation symbol streams.

19. In a multiple-input multiple-output (MIMO) communication system, a method for processing data for transmission over a MIMO channel, comprising:

processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

obtaining an estimated channel response matrix for the MIMO channel;

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors and a plurality of matrices of singular values;

deriving a plurality of weights based on the matrices of singular values, wherein the weights are used to invert the frequency response of the MIMO channel;

deriving a pulse-shaping matrix based on the matrices of eigen-vectors and the weights; and

preconditioning the plurality of streams of modulation symbols based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

20. The method of claim 19, further comprising:

deriving a plurality of scaling values based on the matrices of singular values, wherein the scaling values are used to adjust transmit powers for eigenmodes of the

MIMO channel, and wherein the pulse-shaping matrix is further derived based on the scaling values.

21. A memory communicatively coupled to a digital signal processing device (DSPD) capable of interpreting digital information to:

process data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

derive a pulse-shaping matrix based on an estimated response of the MIMO channel and in a manner to reduce intersymbol interference at a receiver; and

precondition the plurality of streams of modulation symbols based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

22. In a multiple-input multiple-output (MIMO) communication system, a method for processing a data transmission received via a MIMO channel, comprising:

obtaining an estimated channel response matrix for the MIMO channel;

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

deriving a pulse-shaping matrix based on the matrices of eigen-vectors; and

conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

23. The method of claim 22, wherein the conditioning is performed in the time domain based on a time-domain pulse-shaping matrix.

24. The method of claim 22, wherein the conditioning is performed in the frequency domain and includes

transforming the plurality of received symbol streams to the frequency domain;

multiplying the transformed received symbol streams with a frequency-domain pulse-shaping matrix to provide a plurality of conditioned symbol streams; and

transforming the plurality of conditioned symbol streams to the time domain to provide the plurality of recovered symbol streams.

25. The method of claim 22, wherein the conditioning orthogonalizes a plurality of streams of modulation symbols transmitted over the MIMO channel.

26. The method of claim 22, further comprising:

demodulating the plurality of recovered symbol streams in accordance with one or more demodulation schemes to provide a plurality of demodulated data streams; and

decoding the plurality of demodulated data streams in accordance with one or more decoding schemes to provide decoded data.

27. The method of claim 22, further comprising:

deriving channel state information (CSI) comprised of the estimated channel response matrix for the MIMO channel; and

sending the CSI back to the transmitter.

28. In a multiple-input multiple-output (MIMO) communication system, a method for processing a data transmission received via a MIMO channel, comprising:

obtaining an estimated channel response matrix for the MIMO channel;

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

deriving a pulse-shaping matrix based on the matrices of eigen-vectors;

conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver;

demodulating the plurality of recovered symbol streams in accordance with one or more demodulation schemes to provide a plurality of demodulated data streams; and

decoding the plurality of demodulated data streams in accordance with one or more decoding schemes to provide decoded data.

29. A memory communicatively coupled to a digital signal processing device (DSPD) capable of interpreting digital information to:

obtain an estimated channel response matrix for a MIMO channel used for a data transmission;

decompose the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

derive a pulse-shaping matrix based on the matrices of eigen-vectors; and

condition a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

30. A transmitter unit in a multiple-input multiple-output (MIMO) communication system, comprising:

a TX data processor operative to process data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols; and

a TX MIMO processor operative to derive a pulse-shaping matrix based on an estimated response of a MIMO channel and in a manner to reduce intersymbol interference at a receiver, and to precondition the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

31. The transmitter unit of claim 30, wherein the TX MIMO processor is further operative to derive a plurality of weights based on an estimated channel response matrix for the MIMO channel, wherein the weights are used to invert the frequency response of the MIMO channel, and wherein the pulse-shaping matrix is derived based in part on the weights.

32. The transmitter unit of claim 31, wherein the TX MIMO processor is further operative to decompose the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors and a plurality of matrices of singular values, and

wherein the weights are derived based on the matrices of singular values and the pulse-shaping matrix is further derived based on the matrices of eigen-vectors.

33. The transmitter unit of claim 31, wherein the TX MIMO processor is further operative to derive a plurality of scaling values used to adjust transmit powers for the eigenmodes of the MIMO channel, and wherein the pulse-shaping matrix is further derived based on the scaling values.

34. The transmitter unit of claim 33, wherein the scaling values are derived based on water-pouring analysis on a plurality of matrices of singular values obtained from the estimated channel response matrix.

35. An apparatus in a multiple-input multiple-output (MIMO) communication system, comprising:

means for processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

means for deriving a pulse-shaping matrix based on an estimated response of a MIMO channel and in a manner to reduce intersymbol interference at a receiver; and

means for preconditioning the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

36. A digital signal processor comprising:

means for processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

means for deriving a pulse-shaping matrix based on an estimated response of a multiple-input multiple-output (MIMO) channel and in a manner to reduce intersymbol interference at a receiver; and

means for preconditioning the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

37. A receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

an RX MIMO processor operative to obtain an estimated channel response matrix for a MIMO channel used for a data transmission, decompose the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors, derive a pulse-shaping matrix based on the matrices of eigen-vectors, and condition a plurality of streams of received symbols based on the pulse-shaping matrix to obtain a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted over the MIMO channel, wherein the modulation symbols were preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at the receiver unit; and

an RX data processor operative to process the plurality of recovered symbol streams in accordance with a particular processing scheme to provide decoded data.

38. The receiver unit of claim 37, wherein the RX MIMO processor is operative to condition the plurality of streams of received symbols in the time domain based on a time-domain pulse-shaping matrix.

39. An apparatus in a multiple-input multiple-output (MIMO) communication system, comprising:

means for obtaining an estimated channel response matrix for a MIMO channel used for a data transmission;

means for decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

means for deriving a pulse-shaping matrix based on the matrices of eigen-vectors; and

means for conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

40. A digital signal processor comprising:

means for obtaining an estimated channel response matrix for a multiple-input multiple-output (MIMO) channel used for a data transmission;

means for decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

means for deriving a pulse-shaping matrix based on the matrices of eigen-vectors; and

means for conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

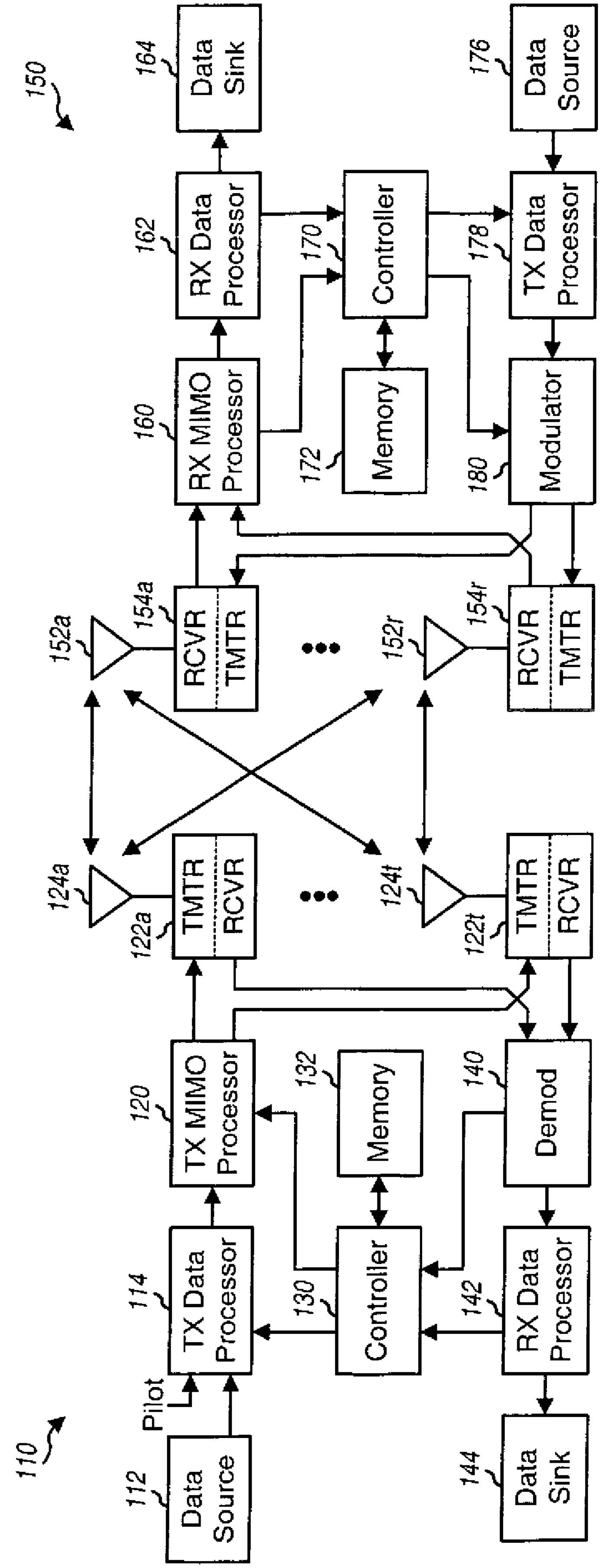
150
164
Data Sink
176
Data Source
162
RX Data Processor
170
Controller
178
TX Data Processor
160
RX MIMO Processor
172
Memory
180
Modulator
154a
RCVR
TMTR
152a
152r
154r
RCVR
TMTR
124a
TMTR
RCVR
122a
124t
122t
TMTR
RCVR
120
TX MIMO Processor
132
Memory
140
Demod
114
TX Data Processor
130
Controller
142
RX Data Processor
Pilot
110
112
Data Source
144
Data Sink

FIG. 1

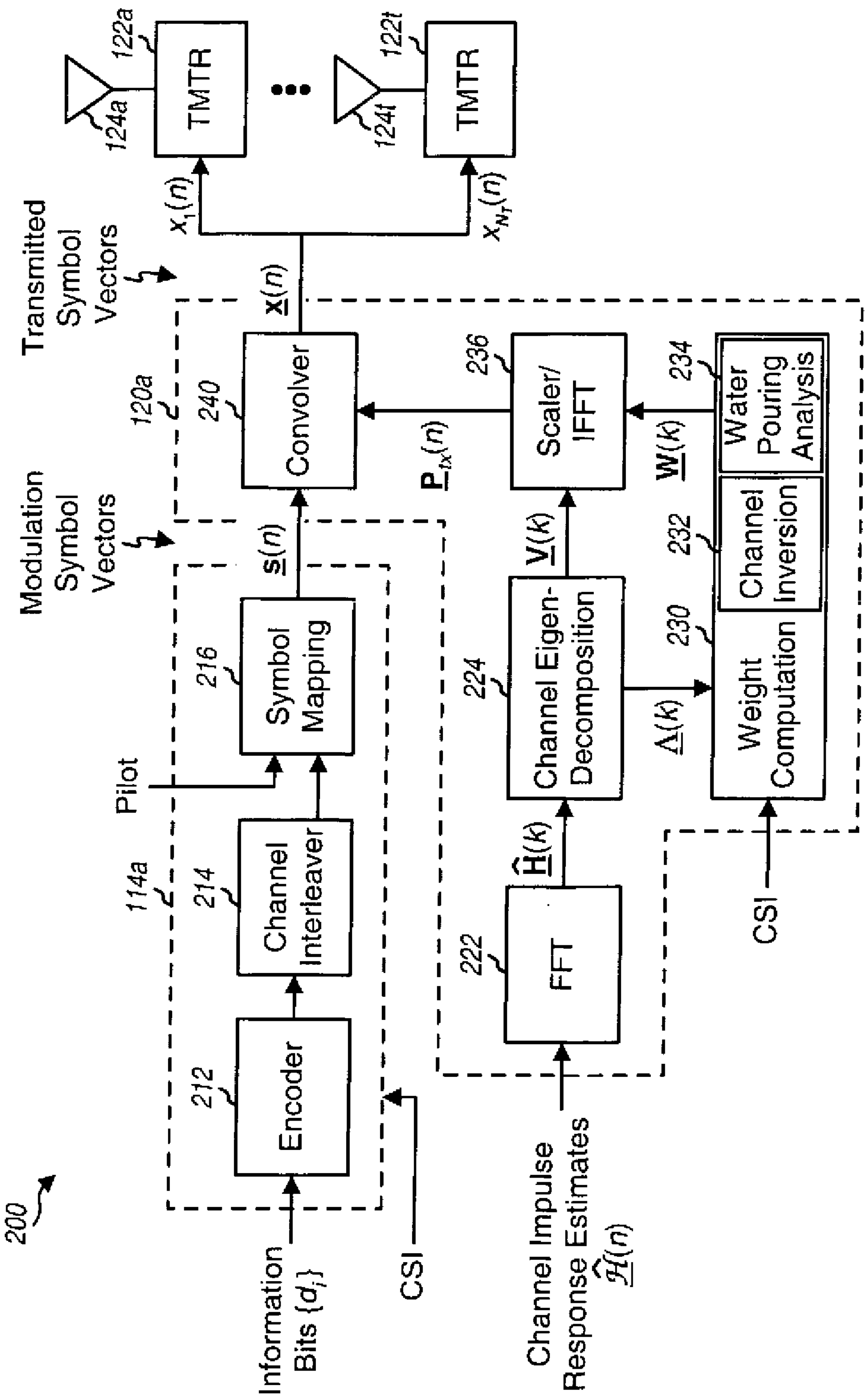
200
Information Bits $\{d_i\}$
CSI
Encoder
212
Channel Interleaver
214
114a
Pilot
Symbol Mapping
216
Modulation Symbol Vectors
$\underline{s}(n)$
Convolver
240
120a
Transmitted Symbol Vectors
$\underline{x}(n)$
$x_1(n)$
$x_{N_T}(n)$
TMTR
122a
124a
TMTR
122t
124t
Channel Impulse Response Estimates $\hat{\mathcal{H}}(n)$
FFT
222
$\hat{\underline{H}}(k)$
Channel Eigen-Decomposition
224
$\underline{V}(k)$
Scaler/ IFFT
236
$\underline{P}_{tx}(n)$
$\underline{\Lambda}(k)$
Weight Computation
230
Channel Inversion
232
Water Pouring Analysis
234
$\underline{W}(k)$
CSI

FIG. 2

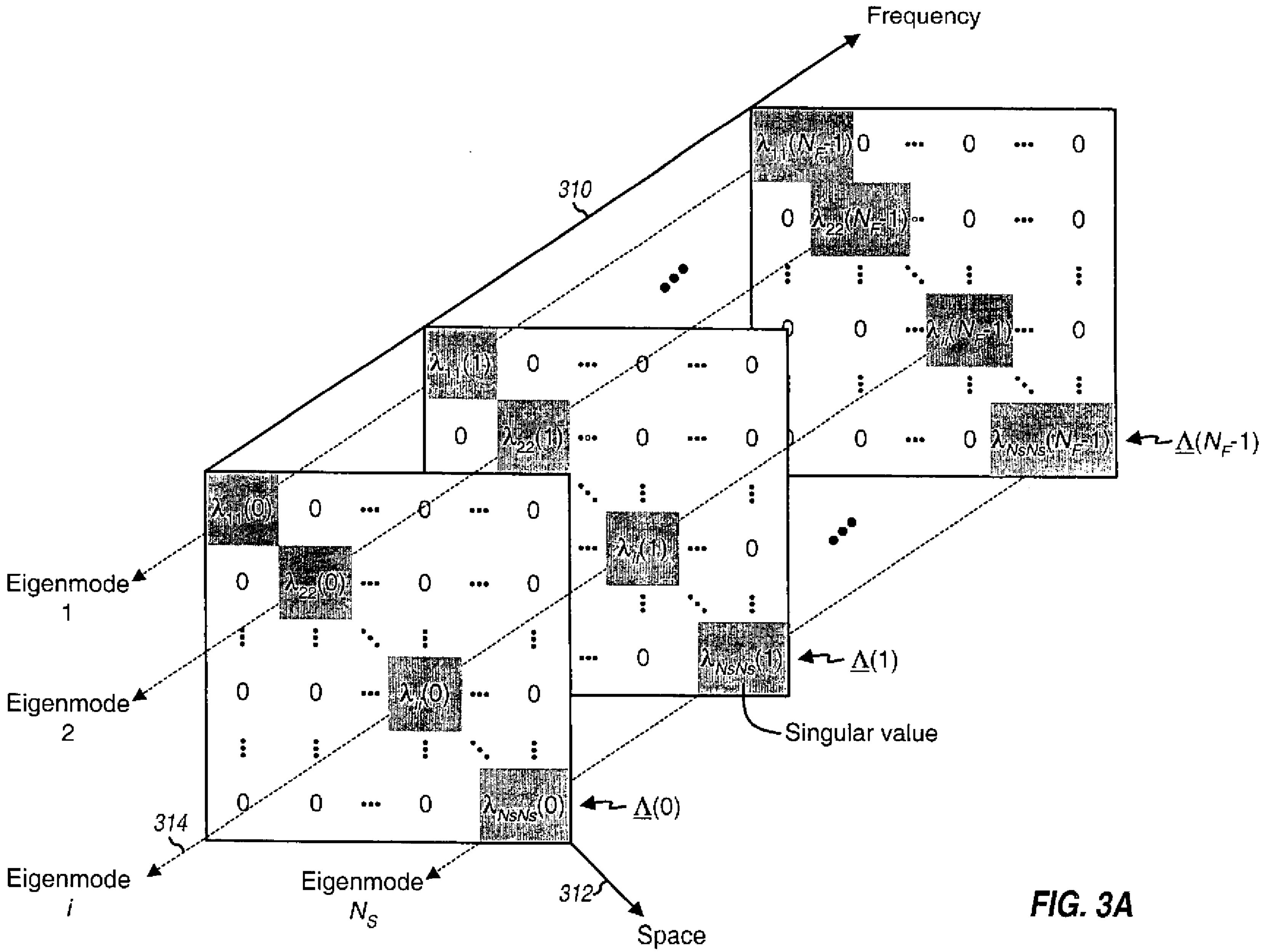
Frequency
310
$\underline{\Lambda}(N_F-1)$
$\underline{\Lambda}(1)$
Singular value
$\underline{\Lambda}(0)$
Eigenmode 1
Eigenmode 2
314
Eigenmode i
Eigenmode N_S
312
Space

FIG. 3A

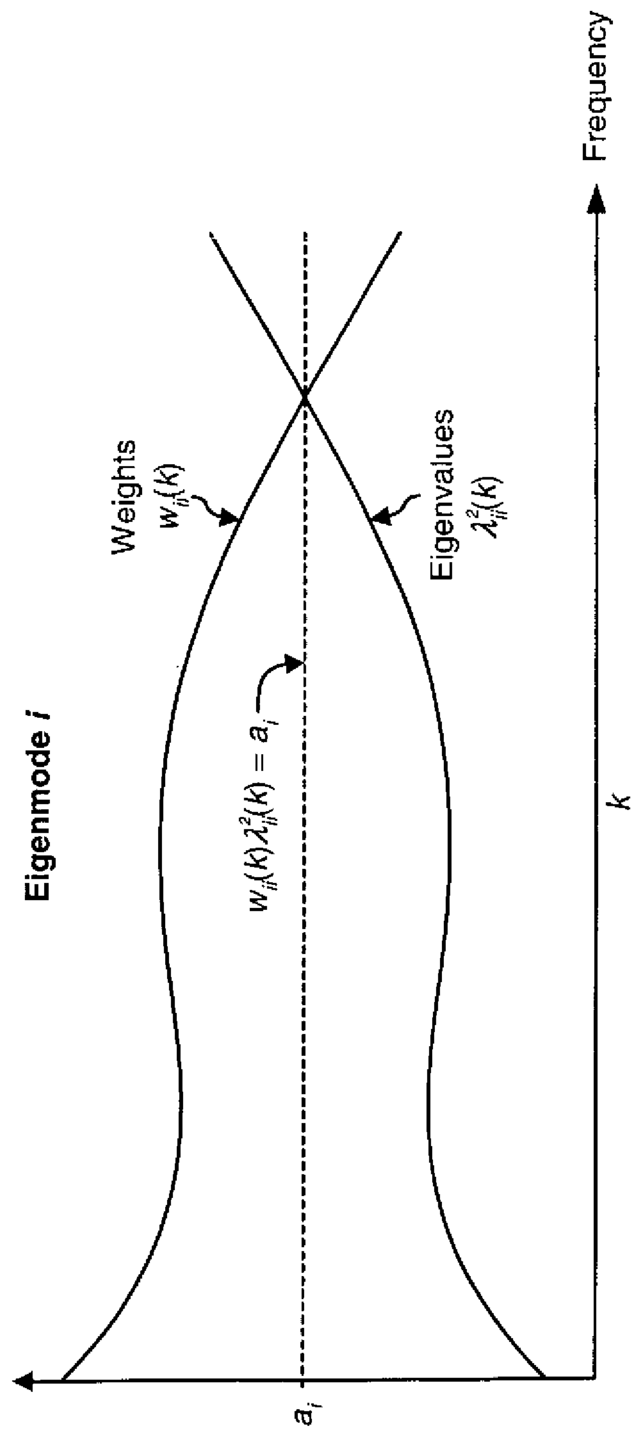
Eigenmode *i*
Weights
$w_{ii}(k)$
Eigenvalues
$\lambda^2_{ii}(k)$
$w_{ii}(k)\lambda^2_{ii}(k) = a_i$
a_i
k
Frequency

FIG. 3B

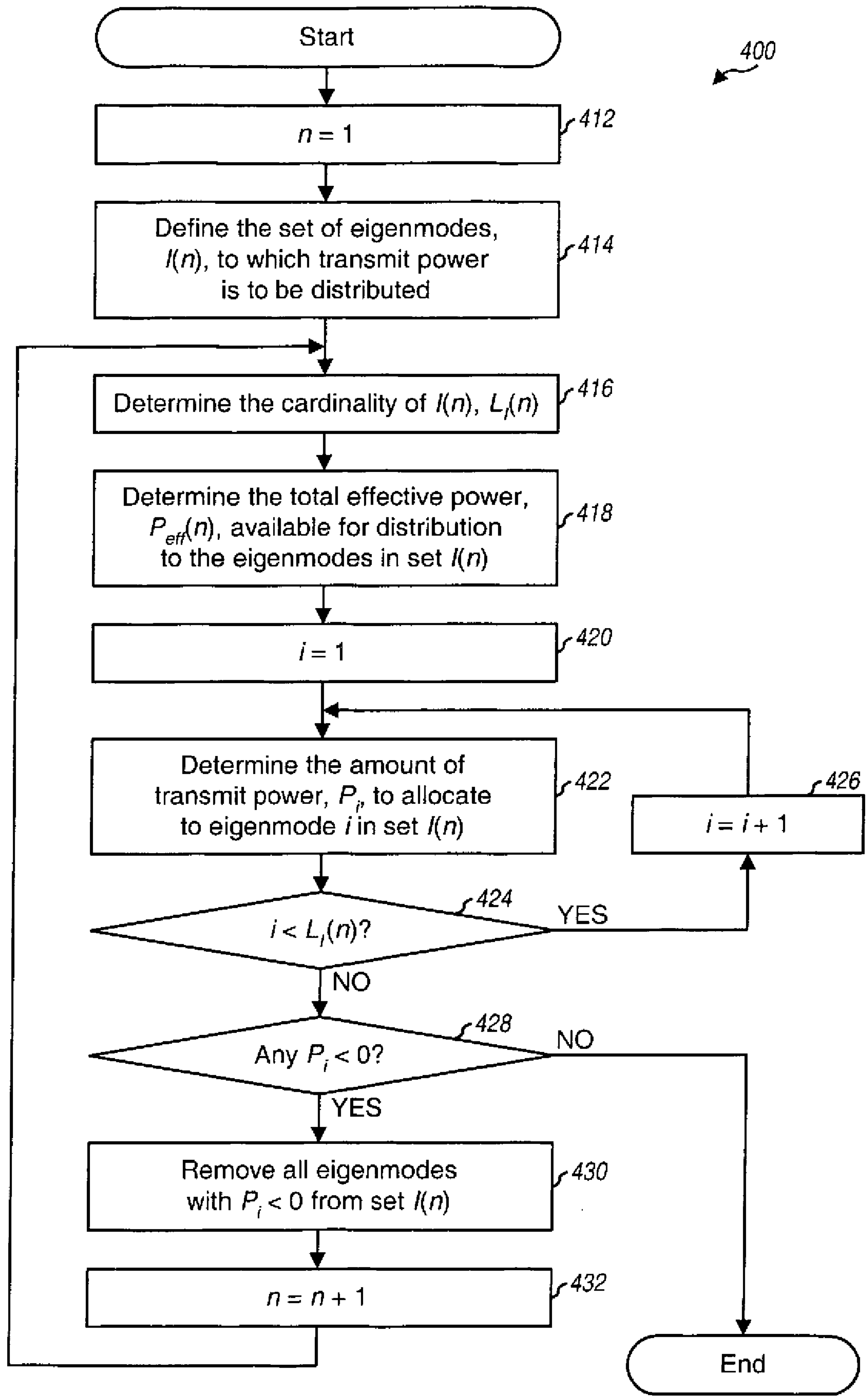
Start
400
$n = 1$
412
Define the set of eigenmodes, $I(n)$, to which transmit power is to be distributed
414
Determine the cardinality of $I(n)$, $L_I(n)$
416
Determine the total effective power, $P_{eff}(n)$, available for distribution to the eigenmodes in set $I(n)$
418
$i = 1$
420
Determine the amount of transmit power, P_i, to allocate to eigenmode i in set $I(n)$
422
$i = i + 1$
426
$i < L_I(n)$?
424
YES
NO
Any $P_i < 0$?
428
NO
YES
Remove all eigenmodes with $P_i < 0$ from set $I(n)$
430
$n = n + 1$
432
End

FIG. 4

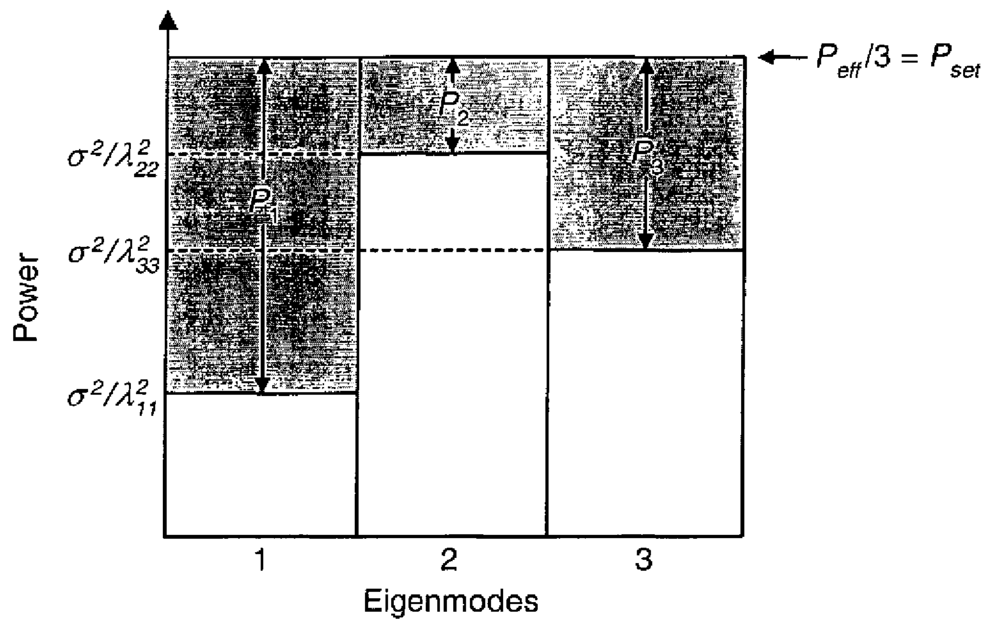
$P_{eff}/3 = P_{set}$
σ^2/λ_{22}^2
σ^2/λ_{33}^2
σ^2/λ_{11}^2
P_1
P_2
P_3
Power
1
2
3
Eigenmodes

FIG. 5A

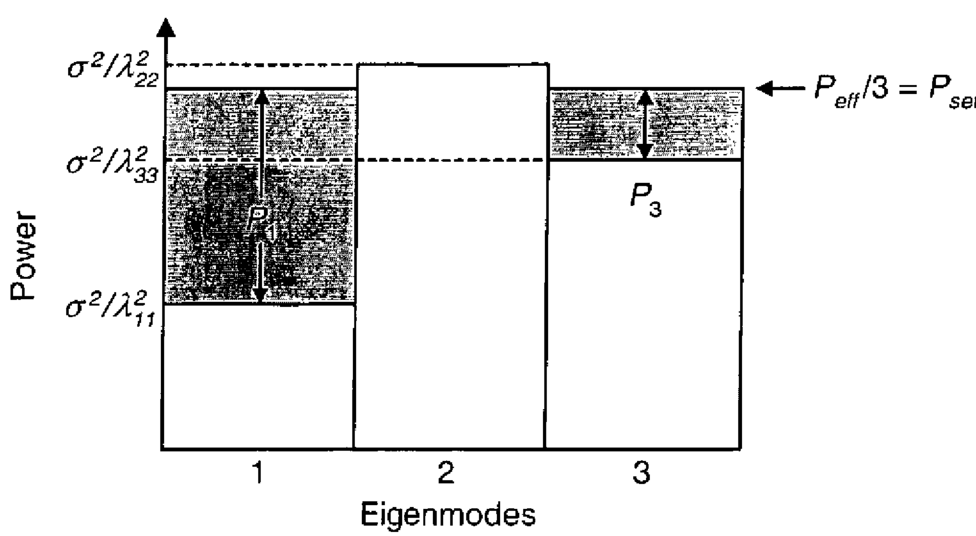
σ^2/λ_{22}^2
$P_{eff}/3 = P_{set}$
σ^2/λ_{33}^2
σ^2/λ_{11}^2
P_1
P_3
Power
1
2
3
Eigenmodes

FIG. 5B

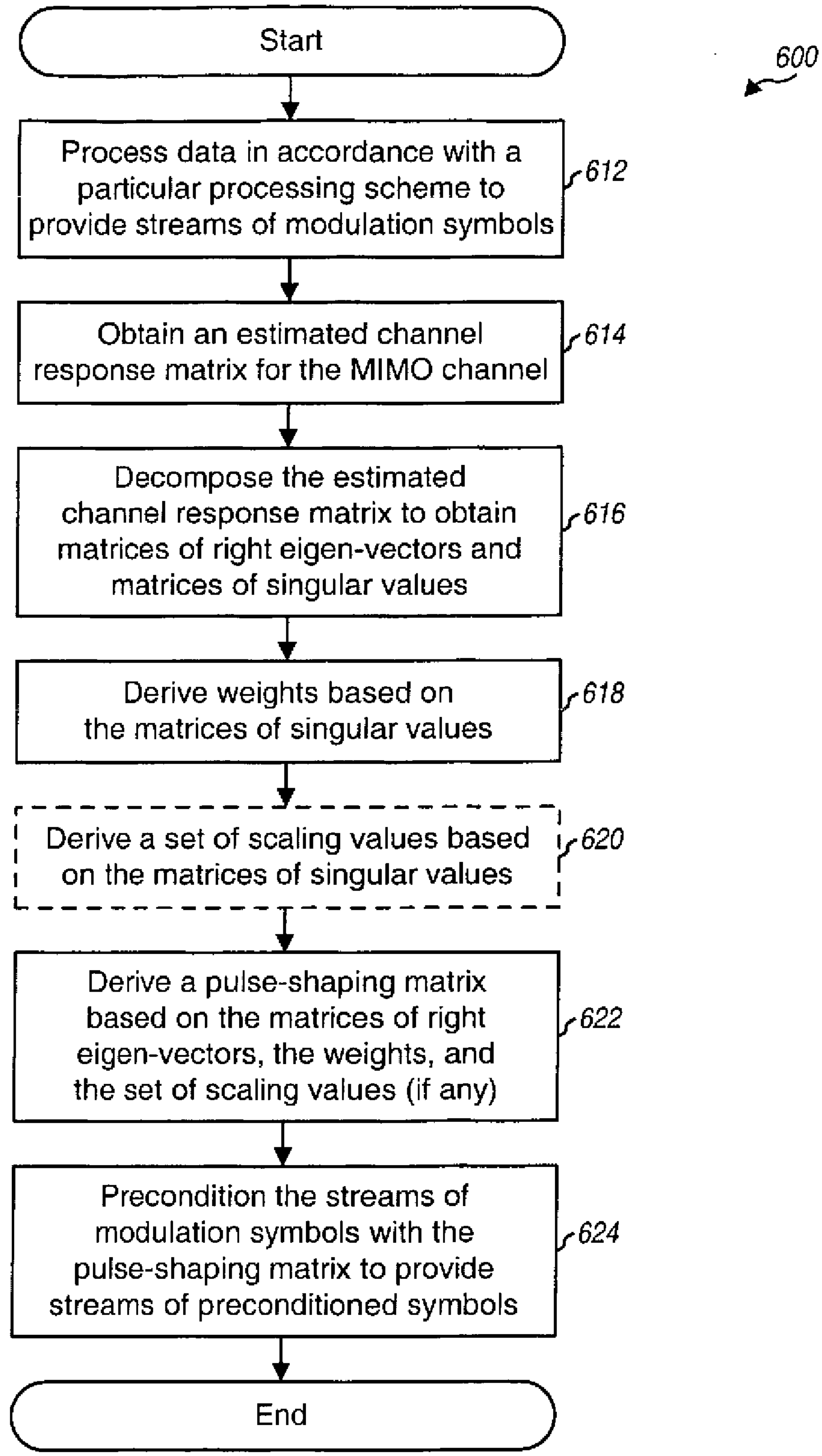
Start
600
Process data in accordance with a particular processing scheme to provide streams of modulation symbols
612
Obtain an estimated channel response matrix for the MIMO channel
614
Decompose the estimated channel response matrix to obtain matrices of right eigen-vectors and matrices of singular values
616
Derive weights based on the matrices of singular values
618
Derive a set of scaling values based on the matrices of singular values
620
Derive a pulse-shaping matrix based on the matrices of right eigen-vectors, the weights, and the set of scaling values (if any)
622
Precondition the streams of modulation symbols with the pulse-shaping matrix to provide streams of preconditioned symbols
624
End

FIG. 6

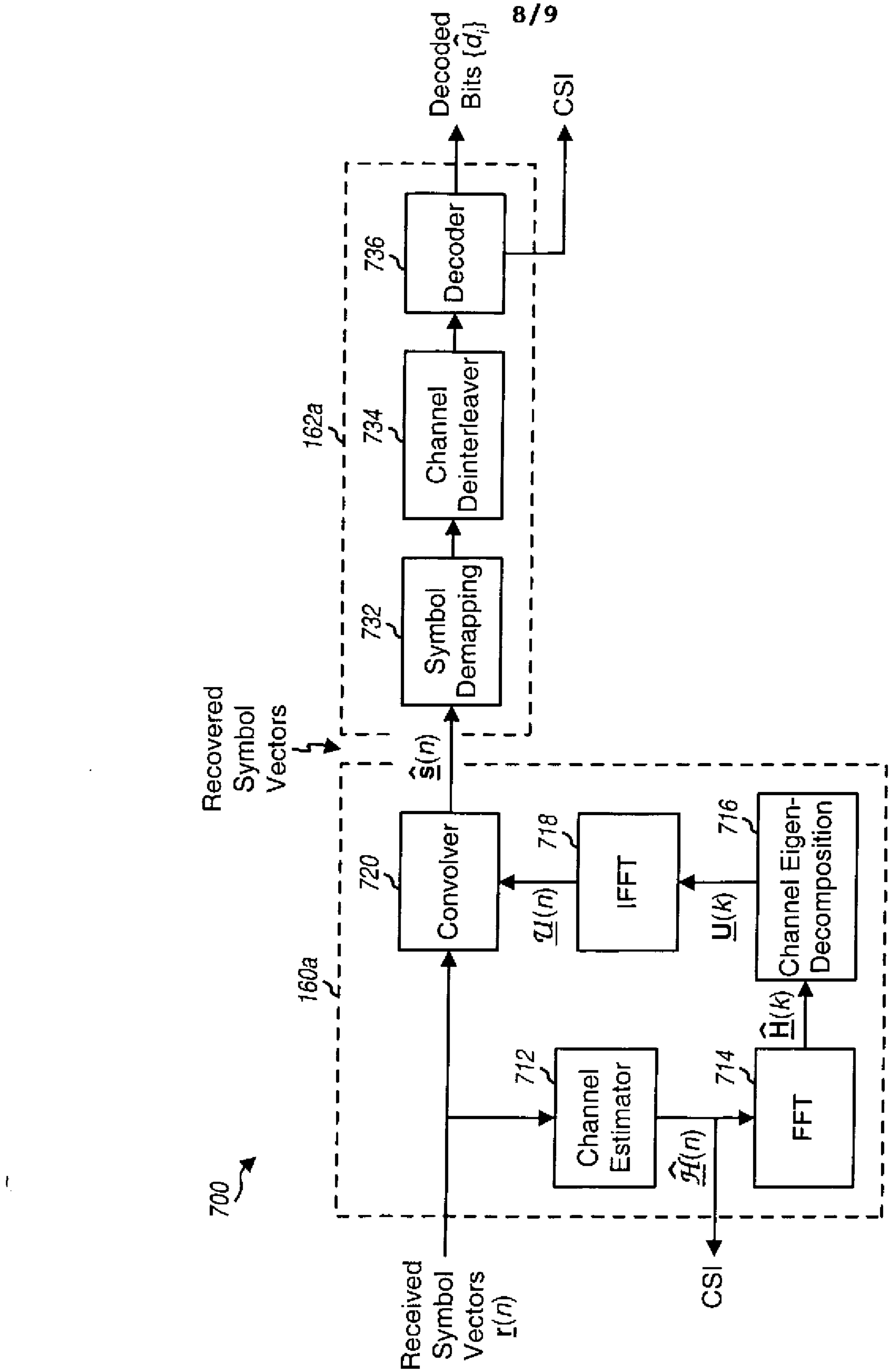
700
Received Symbol Vectors $\underline{r}(n)$
160a
Channel Estimator
712
$\hat{\underline{\mathcal{H}}}(n)$
CSI
FFT
714
$\hat{\underline{H}}(k)$
Channel Eigen-Decomposition
716
$\underline{U}(k)$
IFFT
718
$\underline{\mathcal{U}}(n)$
Convolver
720
$\hat{\underline{s}}(n)$
Recovered Symbol Vectors
162a
Symbol Demapping
732
Channel Deinterleaver
734
Decoder
736
Decoded Bits $\{\hat{d}_i\}$
CSI

FIG. 7

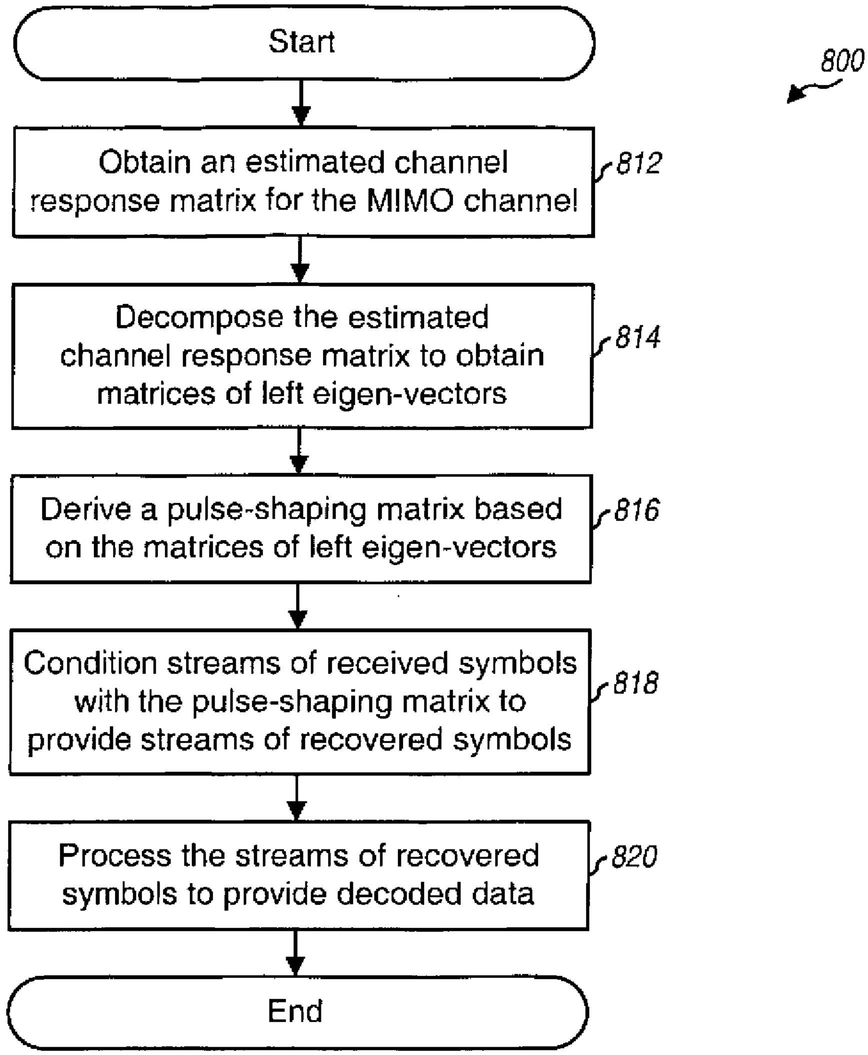
800
Start
Obtain an estimated channel response matrix for the MIMO channel
812
Decompose the estimated channel response matrix to obtain matrices of left eigen-vectors
814
Derive a pulse-shaping matrix based on the matrices of left eigen-vectors
816
Condition streams of received symbols with the pulse-shaping matrix to provide streams of recovered symbols
818
Process the streams of recovered symbols to provide decoded data
820
End

FIG. 8

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 03/19464

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L1/06 H04L25/03 H04L25/02

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, PAJ, INSPEC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	SAMPATH H ET AL: "Joint transmit and receive optimization for high data rate wireless communication using multiple antennas" SIGNALS, SYSTEMS, AND COMPUTERS, 1999. CONFERENCE RECORD OF THE THIRTY-THIRD ASILOMAR CONFERENCE ON OCT. 24-27, 1999, PISCATAWAY, NJ, USA,IEEE, US, 24 October 1999 (1999-10-24), pages 215-219, XP010373976 ISBN: 0-7803-5700-0 abstract page 215, paragraph 2 page 215, paragraph 5 -page 216, paragraph 2 page 217, column 2 -page 218, paragraph 1	1,13-18, 21-30, 35-40
A	---	2-12,19, 20,31-34

-/--

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
24 September 2003	06/10/2003

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Reilly, D

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	WO 02 05506 A (QUALCOMM INC) 17 January 2002 (2002-01-17) abstract page 13, line 4 - line 11 page 14, line 24 -page 19, line 4	1,13-18, 21-30, 35-40
A		2-12,19, 20,31-34

X	JOONSUK KIM ET AL: "Transmission Optimization with a Space-Time Filter at Low SNR Wireless Environment" GLOBECOM'99, vol. 1B, 5 December 1999 (1999-12-05), pages 889-893, XP010373673 abstract page 890, column 1, last paragraph -page 892, column 1, paragraph 2	1,13-18, 21-30, 35-40
A		2-12,19, 20,31-34

X	BURR A G: "Adaptive space-time signal processing and coding" IEEE, vol. 2, 22 October 2000 (2000-10-22), pages 710-714, XP010531996 abstract page 710, column 2, paragraph 3 -page 712, column 1, paragraph 3	1,13-18, 21-30, 35-40
A		2-12,19, 20,31-34

A	EP 0 549 019 A (KONINKL PHILIPS ELECTRONICS NV) 30 June 1993 (1993-06-30) abstract page 2, line 25 - line 30	2-12,19, 20,31-34

INTERNATIONAL SEARCH REPORT

Information on patent family members

International Application No
PCT/US 03/19464

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
WO 0205506 A	17-01-2002	AU 7341101 A	21-01-2002
		EP 1299978 A2	09-04-2003
		WO 0205506 A2	17-01-2002
EP 0549019 A	30-06-1993	DE 69227356 D1	26-11-1998
		DE 69227356 T2	27-05-1999
		EP 0549019 A2	30-06-1993
		JP 5327558 A	10-12-1993
		KR 260809 B1	01-07-2000
		US 5353310 A	04-10-1994

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International Bureau

(43) International Publication Date
22 August 2002 (22.08.2002)

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(10) International Publication Number
WO 02/065664 A2

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(21) International Application Number: PCT/US02/05171

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(25) Filing Language: English

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(30) Priority Data:
09/788,259 15 February 2001 (15.02.2001) US

(71) Applicant: QUALCOMM INCORPORATED [US/US]; 5775 Morehouse Drive, San Diego, CA 92121-1714 (US).

(72) Inventors: TIEDEMANN, Edward, G., Jr.; 656 Barretts Mill Road, Concord, MA 01742 (US). CHEN, Tao; 5415 Harvest Run Drive, San Diego, CA 92130 (US). JAIN, Avinash; 11143 Caminito Alvarez, San Diego, CA 92126 (US).

(74) Agents: WADSWORTH, Philip, R. et al.; Qualcomm Incorporated, 5775 Morehouse Drive, San Diego, CA 92121-1714 (US).

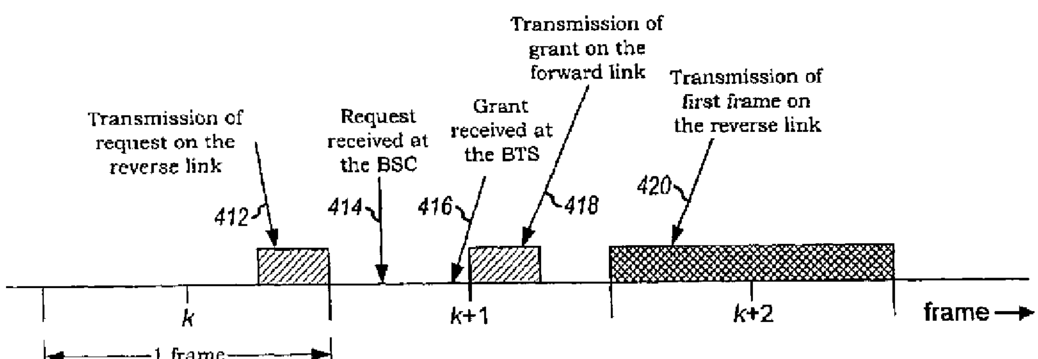
(81) Designated States (national): AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO, NZ, OM, PH, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TN, TR, TT, TZ, UA, UG, UZ, VN, YU, ZA, ZM, ZW.

(84) Designated States (regional): ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:
— *without international search report and to be republished upon receipt of that report*

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

(54) Title: REVERSE LINK CHANNEL ARCHITECTURE FOR A WIRELESS COMMUNICATION SYSTEM



(57) Abstract: A channel structure and mechanisms that support effective and efficient allocation and utilization of the reverse link resources. In one aspect, mechanisms are provided to quickly assign resources (e.g., a supplemental channel) as needed, and to quickly de-assign the resources when not needed or to maintain system stability. The reverse link resources may be quickly assigned and de-assigned via short messages (412, 418) exchanged on control channels on the forward and reverse links. In another aspect, mechanisms are provided to facilitate efficient and reliable data transmission. A reliable acknowledgment/negative acknowledgment scheme and an efficient retransmission scheme are provided. Mechanisms are also provided to control the transmit power and/or data rate of the remote terminals to achieve high performance and avoid instability.

WO 02/065664 A2

REVERSE LINK CHANNEL ARCHITECTURE FOR A WIRELESS COMMUNICATION SYSTEM

BACKGROUND

Field

[1001] The present invention relates generally to data communication, and more specifically to a novel and improved reverse link architecture for a wireless communication system.

Background

[1002] Wireless communication systems are widely deployed to provide various types of communication including voice and packet data services. These systems may be based on code division multiple access (CDMA), time division multiple access (TDMA), or some other modulation techniques. CDMA systems may provide certain advantages over other types of system, including increased system capacity.

[1003] In a wireless communication system, a user with a remote terminal (e.g., a cellular phone) communicates with another user through transmissions on the forward and reverse links via one or more base stations. The forward link (i.e., downlink) refers to transmission from the base station to the user terminal, and the reverse link (i.e., uplink) refers to transmission from the user terminal to the base station. The forward and reverse links are typically allocated different frequencies, a method called frequency division multiplexing (FDM).

[1004] The characteristics of packet data transmission on the forward and reverse links are typically very different. On the forward link, the base station usually knows whether or not it has data to transmit, the amount of data, and the identity of the recipient remote terminals. The base station may further be provided with the "efficiency" achieved by each recipient remote terminal, which may be quantified as the amount of transmit power needed per bit. Based on the known information, the base station may be able to efficiently schedule data

transmissions to the remote terminals at the times and data rates selected to achieve the desired performance.

[1005] On the reverse link, the base station typically does not know *a priori* which remote terminals have packet data to transmit, or how much. The base station is typically aware of each received remote terminal's efficiency, which may be quantified by the energy-per-bit-to-total-noise-plus-interface ratio, Ec/(No+Io), needed at the base station to correctly receive a data transmission. The base station may then allocate resources to the remote terminals whenever requested and as available.

[1006] Because of uncertainty in user demands, the usage on the reverse link may fluctuate widely. If many remote terminals transmit at the same time, high interference is generated at the base station. The transmit power from the remote terminals would need to be increased to maintain the target Ec/(No+Io), which would then result in higher levels of interference. If the transmit power is further increased in this manner, a "black out" may ultimately result and the transmissions from all or a large percentage of the remote terminals may not be properly received. This is due to the remote terminal not being able to transmit at sufficient power to close the link to the base station.

[1007] In a CDMA system, the channel loading on the reverse link is often characterized by what is referred to as the "rise-over-thermal". The rise-over-thermal is the ratio of the total received power at a base station receiver to the power of the thermal noise. Based on theoretical capacity calculations for a CDMA reverse link, there is a theoretical curve that shows the rise-over-thermal increasing with loading. The loading at which the rise-over-thermal is infinite is often referred to as the "pole". A loading that has a rise-over-thermal of 3 dB corresponds to a loading of about 50%, or about half of the number of users that can be supported when at the pole. As the number of users increases and as the data rates of the users increase, the loading becomes higher. Correspondingly, as the loading increases, the amount of power that a remote terminal must transmit increases. The rise-over-thermal and channel loading are described in further detail by A.J. Viterbi in "CDMA : Principles of Spread Spectrum Communication," Addison-Wesley Wireless Communications Series, May 1995, ISBN: 0201633744, which is incorporated herein by reference.

[1008] The Viterbi reference provides classical equations that show the relationship between the rise-over-thermal, the number of users, and the data rates of the users. The equations also show that there is greater capacity (in bits/second) if a few users transmit at a high rate than a larger number of users transmit at a higher rate. This is due to the interference between transmitting users.

[1009] In a typical CDMA system, many users' data rates are continuously changing. For example, in an IS-95 or cdma2000 system, a voice user typically transmits at one of four rates, corresponding to the voice activity at the remote terminal, as described in U.S Patent Nos. 5,657,420 and 5,778,338, both entitled "VARIABLE RATE VOCODER" and U.S Patent No. 5,742,734, entitled "ENCODING RATE SELECTION IN A VARIABLE RATE VOCODER". Similarly, many data users are continually varying their data rates. All this creates a considerable amount of variation in the amount of data being transmitted simultaneously, and hence a considerable variation in the rise-over-thermal.

[1010] As can be seen from the above, there is a need in the art for a reverse link channel structure capable of achieving high performance for packet data transmission, and which takes into consideration the data transmission characteristics of the reverse links.

SUMMARY

[1011] Aspects of the invention provide mechanisms that support effective and efficient allocation and utilization of the reverse link resources. In one aspect, mechanisms are provided to quickly assign resources (e.g., supplemental channels) as needed, and to quickly de-assign the resources when not needed or to maintain system stability. The reverse link resources may be quickly assigned and de-assigned via short messages exchanged on control channels on the forward and reverse links. In another aspect, mechanisms are provided to facilitate efficient and reliable data transmission. In particular, a reliable acknowledgment/negative acknowledgment scheme and an efficient retransmission scheme are provided. In yet another aspect,

mechanisms are provided to control the transmit power and/or data rate of the remote terminals to achieve high performance and avoid instability. Another aspect of the invention provides a channel structure capable of implementing the features described above. These and other aspects are described in further detail below.

[1012] The disclosed embodiments further provide methods, channel structures, and apparatus that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1013] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1014] FIG. 1 is a diagram of a wireless communication system that supports a number of users;

[1015] FIG. 2 is a simplified block diagram of an embodiment of a base station and a remote terminal;

[1016] FIGS. 3A and 3B are diagrams of a reverse and a forward channel structure, respectively;

[1017] FIG. 4 is a diagram illustrating a communication between the remote terminal and base station to assign a reverse link supplemental channel (R-SCH);

[1018] FIGS. 5A and 5B are diagrams illustrating a data transmission on the reverse link and an Ack/Nak message transmission for two different scenarios;

[1019] FIGS. 6A and 6B are diagrams illustrating an acknowledgment sequencing with short and long acknowledgment delays, respectively;

[1020] FIG. 7 is a flow diagram that illustrates a variable rate data transmission on the R-SCH with fast congestion control, in accordance with an embodiment of the invention; and

[1021] FIG. 8 is a diagram illustrating improvement that may be possible with fast control of the R-SCH.

DETAILED DESCRIPTION

[1022] FIG. 1 is a diagram of a wireless communication system 100 that supports a number of users and capable of implementing various aspects of the invention. System 100 provides communication for a number of cells, with each cell being serviced by a corresponding base station 104. The base stations are also commonly referred to as base transceiver systems (BTSs). Various remote terminals 106 are dispersed throughout the system. Each remote terminal 106 may communicate with one or more base stations 104 on the forward and reverse links at any particular moment, depending on whether or not the remote terminal is active and whether or not it is in soft handoff. The forward link refers to transmission from base station 104 to remote terminal 106, and the reverse link refers to transmission from remote terminal 106 to base station 104. As shown in FIG. 1, base station 104a communicates with remote terminals 106a, 106b, 106c, and 106d, and base station 104b communicates with remote terminals 106d, 106e, and 106f. Remote terminal 106d is in soft handoff and concurrently communicates with base stations 104a and 104b.

[1023] In system 100, a base station controller (BSC) 102 couples to base stations 104 and may further couple to a public switched telephone network (PSTN). The coupling to the PSTN is typically achieved via a mobile switching center (MSC), which is not shown in FIG. 1 for simplicity. The BSC may also couple into a packet network, which is typically achieved via a packet data serving node (PDSN) that is also not shown in FIG. 1. BSC 102 provides coordination and control for the base stations coupled to it. BSC 102 further controls the routing of telephone calls among remote terminals 106, and between remote terminals 106 and users coupled to the PSTN (e.g., conventional telephones) and to the packet network, via base stations 104.

[1024] System 100 may be designed to support one or more CDMA standards such as (1) the "TIA/EIA-95-B Mobile Station-Base Station Compatibility Standard for Dual-Mode Wideband Spread Spectrum Cellular System" (the IS-95 standard), (2) the "TIA/EIA-98-D Recommended Minimum Standard for Dual-Mode Wideband Spread Spectrum Cellular Mobile Station"

(the IS-98 standard), (3) the documents offered by a consortium named "3rd Generation Partnership Project" (3GPP) and embodied in a set of documents including Document Nos. 3G TS 25.211, 3G TS 25.212, 3G TS 25.213, and 3G TS 25.214 (the W-CDMA standard), (4) the documents offered by a consortium named "3rd Generation Partnership Project 2" (3GPP2) and embodied in a set of documents including Document Nos. C.S0002-A, C.S0005-A, C.S0010-A, C.S0011-A. C.S0024, and C.S0026 (the cdma2000 standard), and (5) some other standards. In the case of the 3GPP and 3GPP2 documents, these are converted by standards bodies worldwide (e.g., TIA, ETSI, ARIB, TTA, and CWTS) into regional standards and have been converted into international standards by the International Telecommunications Union (ITU). These standards are incorporated herein by reference.

[1025] FIG. 2 is a simplified block diagram of an embodiment of base station 104 and remote terminal 106, which are capable of implementing various aspects of the invention. For a particular communication, voice data, packet data, and/or messages may be exchanged between base station 104 and remote terminal 106. Various types of messages may be transmitted such as messages used to establish a communication session between the base station and remote terminal and messages used to control a data transmission (e.g., power control, data rate information, acknowledgment, and so on). Some of these message types are described in further detail below.

[1026] For the reverse link, at remote terminal 106, voice and/or packet data (e.g., from a data source 210) and messages (e.g., from a controller 230) are provided to a transmit (TX) data processor 212, which formats and encodes the data and messages with one or more coding schemes to generate coded data. Each coding scheme may include any combination of cyclic redundancy check (CRC), convolutional, Turbo, block, and other coding, or no coding at all. Typically, voice data, packet data, and messages are coded using different schemes, and different types of message may also be coded differently.

[1027] The coded data is then provided to a modulator (MOD) 214 and further processed (e.g., covered, spread with short PN sequences, and scrambled with a long PN sequence assigned to the user terminal). The modulated data is then provided to a transmitter unit (TMTR) 216 and

conditioned (e.g., converted to one or more analog signals, amplified, filtered, and quadrature modulated) to generate a reverse link signal. The reverse link signal is routed through a duplexer (D) 218 and transmitted via an antenna 220 to base station 104.

[1028] At base station 104, the reverse link signal is received by an antenna 250, routed through a duplexer 252, and provided to a receiver unit (RCVR) 254. Receiver unit 254 conditions (e.g., filters, amplifies, downconverts, and digitizes) the received signal and provides samples. A demodulator (DEMOD) 256 receives and processes (e.g., despreads, decovers, and pilot demodulates) the samples to provide recovered symbols. Demodulator 256 may implement a rake receiver that processes multiple instances of the received signal and generates combined symbols. A receive (RX) data processor 258 then decodes the symbols to recover the data and messages transmitted on the reverse link. The recovered voice/packet data is provided to a data sink 260 and the recovered messages may be provided to a controller 270. The processing by demodulator 256 and RX data processor 258 are complementary to that performed at remote terminal 106. Demodulator 256 and RX data processor 258 may further be operated to process multiple transmissions received via multiple channels, e.g., a reverse fundamental channel (R-FCH) and a reverse supplemental channel (R-SCH). Also, transmissions may be received simultaneously from multiple remote terminals, each of which may be transmitting on a reverse fundamental channel, a reverse supplemental channel, or both.

[1029] On the forward link, at base station 104, voice and/or packet data (e.g., from a data source 262) and messages (e.g., from controller 270) are processed (e.g., formatted and encoded) by a transmit (TX) data processor 264, further processed (e.g., covered and spread) by a modulator (MOD) 266, and conditioned (e.g., converted to analog signals, amplified, filtered, and quadrature modulated) by a transmitter unit (TMTR) 268 to generate a forward link signal. The forward link signal is routed through duplexer 252 and transmitted via antenna 250 to remote terminal 106.

[1030] At remote terminal 106, the forward link signal is received by antenna 220, routed through duplexer 218, and provided to a receiver unit 222.

Receiver unit 222 conditions (e.g., downconverts, filters, amplifies, quadrature demodulates, and digitizes) the received signal and provides samples. The samples are processed (e.g., despreaded, decovered, and pilot demodulated) by a demodulator 224 to provide symbols, and the symbols are further processed (e.g., decoded and checked) by a receive data processor 226 to recover the data and messages transmitted on the forward link. The recovered data is provided to a data sink 228, and the recovered messages may be provided to controller 230.

[1031] The reverse link has some characteristics that are very different from those of the forward link. In particular, the data transmission characteristics, soft handoff behaviors, and fading phenomenon are typically very different between the forward and reverse links.

[1032] As noted above, on the reverse link, the base station typically does not know *a priori* which remote terminals have packet data to transmit, or how much. Thus, the base station may allocate resources to the remote terminals whenever requested and as available. Because of uncertainty in user demands, the usage on the reverse link may fluctuate widely.

[1033] In accordance with aspects of the invention, mechanisms are provided to effectively and efficiently allocate and utilize the reverse link resources. In one aspect, mechanisms are provided to quickly assign resources as needed, and to quickly de-assign resources when not needed or to maintain system stability. The reverse link resources may be assigned via a supplemental channel that is used for packet data transmission. In another aspect, mechanisms are provided to facilitate efficient and reliable data transmission. In particular, a reliable acknowledgment scheme and an efficient retransmission scheme are provided. In yet another aspect, mechanisms are provided to control the transmit power of the remote terminals to achieve high performance and avoid instability. These and other aspects are described in further detail below.

[1034] FIG. 3A is a diagram of an embodiment of a reverse channel structure capable of implementing various aspects of the invention. In this embodiment, the reverse channel structure includes an access channel, an enhanced access channel, a pilot channel (R-PICH), a common control channel (R-CCCH), a

dedicated control channel (R-DCCH), a fundamental channel (R-FCH), supplemental channels (R-SCH), and a reverse rate indicator channel (R-RICH). Different, fewer, and/or additional channels may also be supported and are within the scope of the invention. These channels may be implemented similar to those defined by the cdma2000 standard. Features of some of these channels are described below.

[1035] For each communication (i.e., each call), a specific set of channels that may be used for the communication and their configurations are defined by one of a number of radio configurations (RC). Each RC defines a specific transmission format, which is characterized by various physical layer parameters such as, for example, the transmission rates, modulation characteristics, spreading rate, and so on. The radio configurations may be similar to those defined for the cdma2000 standard.

[1036] The reverse dedicated control channel (R-DCCH) is used to transmit user and signaling information (e.g., control information) to the base station during a communication. The R-DCCH may be implemented similar to the R-DCCH defined in the cdma2000 standard.

[1037] The reverse fundamental channel (R-FCH) is used to transmit user and signaling information (e.g., voice data) to the base station during a communication. The R-FCH may be implemented similar to the R-FCH defined in the cdma2000 standard.

[1038] The reverse supplemental channel (R-SCH) is used to transmit user information (e.g., packet data) to the base station during a communication. The R-SCH is supported by some radio configurations (e.g., RC3 through RC11), and is assigned to the remote terminals as needed and if available. In an embodiment, zero, one, or two supplemental channels (i.e., R-SCH1 and R-SCH2) may be assigned to the remote terminal at any given moment. In an embodiment, the R-SCH supports retransmission at the physical layer, and may utilize different coding schemes for the retransmission. For example, a retransmission may use a code rate of 1/2 for the original transmission. The same rate 1/2 code symbols may be repeated for the retransmission. In an alternative embodiment, the underlying code may be a rate 1/4 code. The original transmission may use 1/2 of the symbols and the retransmission may

use the other half of the symbols. If a third retransmission is done, it can repeat one of the group of symbols, part of each group, a subset of either group, and other possible combinations of symbols.

[1039] R-SCH2 may be used in conjunction with R-SCH1 (e.g., for RC11). In particular, R-SCH2 may be used to provide a different quality of service (QoS). Also, Type II and III hybrid ARQ schemes may be used in conjunction with the R-SCH. Hybrid ARQ schemes are generally described by S.B. Wicker in "Error Control System for Digital Communication and Storage," Prentice-Hall, 1995, Chapter 15, which is incorporated herein by reference. Hybrid ARQ schemes are also described in the cdma2000 standard.

[1040] The reverse rate indicator channel (R-RICH) is used by the remote terminal to provide information pertaining to the (packet) transmission rate on one or more reverse supplemental channels. Table 1 lists the fields for a specific format of the R-RICH. In an embodiment, for each data frame transmission on the R-SCH, the remote terminal sends a reverse rate indicator (RRI) symbol, which indicates the data rate for the data frame. The remote terminal also sends the sequence number of the data frame being transmitted, and whether the data frame is a first transmission or a retransmission. Different, fewer, and/or additional fields may also be used for the R-RICH and are within the scope of the invention. The information in Table 1 is sent by the remote terminal for each data frame transmitted on the supplemental channel (e.g., each 20 msec).

Table 1

Field	Length (bits)
RRI	3
SEQUENCE_NUM	2
RETRAN_NUM	2

[1041] If there are multiple reverse supplemental channels (e.g., R-SCH1 and R-SCH2), then there can be multiple R-RICH channels (e.g., R-RICH1 and R-RICH2), each with the RRI, SEQUENCE_NUM, and RETRAN_NUM fields. Alternatively, the fields for multiple reverse supplemental channels may be

combined into a single R-RICH channel. In a particular embodiment, the RRI field is not used, and fixed transmission rates are used or the base station performs blind rate determination in which the base determines the transmission rate from the data. Blind rate determination may be achieved in a manner described in U.S Patent No. 6,175,590, entitled "METHOD AND APPARATUS FOR DETERMINING THE RATE OF RECEIVED DATA IN A VARIABLE RATE COMMUNICATION SYSTEM," issued January 16, 2001, U.S Patent No. 5,751,725, entitled "METHOD AND APPARATUS FOR DETERMINING THE RATE OF RECEIVED DATA IN A VARIABLE RATE COMMUNICATION SYSTEM," issued May 12, 1998, both of which are assigned to the assignee of the present application and incorporated herein by reference.

[1042] FIG. 3B is a diagram of an embodiment of a forward channel structure capable of supporting various aspects of the invention. In this embodiment, the forward channel structure includes common channels, pilot channels, and dedicated channels. The common channels include a broadcast channel (F-BCCH), a quick paging channel (F-QPCH), a common control channel (F-CCCH), and a common power control channel (F-CPCCH). The pilot channels include a basic pilot channel and an auxiliary pilot channel. And the dedicated channels include a fundamental channel (F-FCH), a supplemental channel (F-SCH), a dedicated auxiliary channel (F-APICH), a dedicated control channel (F-DCCH), and a dedicated packet control channel (F-CPDCCH). Again, different, fewer, and/or additional channels may also be supported and are within the scope of the invention. These channels may be implemented similar to those defined by the cdma2000 standard. Features of some of these channels are described below.

[1043] The forward common power control channel (F-CPCCH) is used by the base station to transmit power control subchannels (e.g., one bit per subchannel) for power control of the R-PICH, R-FCH, R-DCCH, and R-SCH. In an embodiment, upon channel assignment, a remote terminal is assigned a reverse link power control subchannel from one of three sources - the F-DCCH, F-SCH, and F-CPCCH. The F-CPCCH may be assigned if the reverse link power control subchannel is not provided from either the F-DCCH or F-SCH.

[1044] In an embodiment, the available bits in the F-CPCCH may be used to form one or more power control subchannels, which may then be assigned for different uses. For example, a number of power control subchannels may be defined and used for power control of a number of reverse link channels. Power control for multiple channels based on multiple power control subchannels may be implemented as described in U.S. Patent No. 5,991,284, entitled "SUBCHANNEL POWER CONTROL," issued November 23, 1999, assigned to the assignee of the present application and incorporated herein by reference.

[1045] In one specific implementation, an 800 bps power control subchannel controls the power of the reverse pilot channel (R-PICH). All reverse traffic channels (e.g., the R-FCH, R-DCCH, and R-SCH) have their power levels related to the R-PICH by a known relationship, e.g., as described in C.S0002. The ratio between two channels is often referred to as the traffic-to-pilot ratio. The traffic-to-pilot ratio (i.e., the power level of the reverse traffic channel relative to the R-PICH) can be adjusted by messaging from the base station. However, this messaging is slow, so a 100 bits/second (bps) power control subchannel may be defined and used for power control of the R-SCH. In an embodiment, this R-SCH power control subchannel controls the R-SCH relative to the R-PICH. In another embodiment, the R-SCH power control subchannel controls the absolute transmission power of the R-SCH.

[1046] In an aspect of the invention, a "congestion" control subchannel may also be defined for control of the R-SCH, and this congestion control subchannel may be implemented based on the R-SCH power control subchannel or another subchannel.

[1047] Power control for the reverse link is described in further detail below.

[1048] The forward dedicated packet control channel (F-DPCCH) is used to transmit user and signaling information to a specific remote terminal during a communication. The F-DPCCH may be used to control a reverse link packet data transmission. In an embodiment, the F-DPCCH is encoded and interleaved to enhance reliability, and may be implemented similar to the F-DCCH defined by the cdma2000 standard.

[1049] Table 2 lists the fields for a specific format of the F-DPCCH. In an embodiment, the F-DPCCH has a frame size of 48 bits, of which 16 are used for CRC, 8 bits are used for the encoder tail, and 24 bits are available for data and messaging. In an embodiment, the default transmission rate for the F-DPCCH is 9600 bps, in which case a 48-bit frame can be transmitted in 5 msec time interval. In an embodiment, each transmission (i.e., each F-DPCCH frame) is covered with a public long code of the recipient remote terminal to which the frame is targeted. This avoids the need to use an explicit address (hence, the channel is referred to as a "dedicated" channel). However, the F-DPCCH is also "common" since a large number of remote terminals in dedicated channel mode may continually monitor the channel. If a message is directed to a particular remote terminal and is received correctly, then the CRC will check.

Table 2

Field	**Number of Bits / Frame**
Information	24
Frame Quality Indicator	16
Encoder Tail	8

[1050] The F-DPCCH may be used to transmit mini-messages, such as the ones defined by the cdma2000 standard. For example, the F-DPCCH may be used to transmit a *Reverse Supplemental Channel Assignment Mini Message* (RSCAMM) used to grant the F-SCH to the remote terminal.

[1051] The forward common packet Ack/Nak channel (F-CPANCH) is used by the base station to transmit (1) acknowledgments (Ack) and negative acknowledgments (Nak) for a reverse link packet data transmission and (2) other control information. In an embodiment, acknowledgments and negative acknowledgments are transmitted as n-bit Ack/Nak messages, with each message being associated with a corresponding data frame transmitted on the reverse link. In an embodiment, each Ack/Nak message may include 1, 2, 3, or 4 bits (or possible more bits), with the number of bits in the message being dependent on the number of reverse link channels in the service configuration.

The n-bit Ack/Nak message may be block coded to increase reliability or transmitted in the clear.

[1052] In an aspect, to improve reliability, the Ack/Nak message for a particular data frame is retransmitted in a subsequent frame (e.g., 20 msec later) to provide time diversity for the message. The time diversity provides additional reliability, or may allow for the reduction in power used to send the Ack/Nak message while maintaining the same reliability. The Ack/Nak message may use error correcting coding as is well known in the art. For the retransmission, the Ack/Nak message may repeat the exact same code word or may use incremental redundancy. Transmission and retransmission of the Ack/Nak is described in further detail below.

[1053] Several types of control are used on the forward link to control the reverse link. These include controls for supplemental channel request and grant, Ack/Nak for a reverse link data transmission, power control of the data transmission, and possibly others.

[1054] The reverse link may be operated to maintain the rise-over-thermal at the base station relatively constant as long as there is reverse link data to be transmitted. Transmission on the R-SCH may be allocated in various ways, two of which are described below:

- By infinite allocation. This method is used for real-time traffic that cannot tolerate much delay. The remote terminal is allowed to transmit immediately up to a certain allocated data rate.
- By scheduling. The remote terminal sends an estimate of its buffer size. The base station determines when the remote terminal is allowed to transmit. This method is used for available bit rate traffic. The goal of a scheduler is to limit the number of simultaneous transmissions so that the number of simultaneously transmitting remote terminals is limited, thus reducing the interference between remote terminals.

[1055] Since channel loading can change relatively dramatically, a fast control mechanism may be used to control the transmit power of the R-SCH (e.g., relative to the reverse pilot channel), as described below.

[1056] A communication between the remote terminal and base station to establish a connection may be achieved as follows. Initially, the remote terminal is in a dormant mode or is monitoring the common channels with the slotted timer active (i.e., the remote terminal is monitoring each slot). At a particular time, the remote terminal desires a data transmission and sends a short message to the base station requesting a reconnection of the link. In response, the base station may send a message specifying the parameters to be used for the communication and the configurations of various channels. This information may be sent via an *Extended Channel Assignment Message* (ECAM), a specially defined message, or some other message. This message may specify the following:

- The MAC_ID for each member of the remote terminal's Active Set or a subset of the Active Set. The MAC_ID is later used for addressing on the forward link.
- Whether the R-DCCH or R-FCH is used on the reverse link.
- For the F-CPANCH, the spreading (e.g., Walsh) codes and Active Set to be used. This may be achieved by (1) sending the spreading codes in the ECAM, or (2) transmitting the spreading codes in a broadcast message, which is received by the remote terminal. The spreading codes of neighbor cells may need to be included. If the same spreading codes can be used in neighboring cells, only a single spreading code may need to be sent.
- For the F-CPCCH, the Active Set, the channel identity, and the bit positions. In an embodiment, the MAC_ID may be hashed to the F-CPCCH bit positions to obviate the need to send the actual bit positions or subchannel ID to the remote terminal. This hashing is a pseudo-random method to map a MAC_ID to a subchannel on the F-CPCCH. Since different simultaneous remote terminals are assigned distinct MAC_IDs, the hashing can be such that these MAC_IDs also map to distinct F-CPCCH subchannels. For example, if there are K possible bit positions and N possible MAC_IDs, then $K = \lfloor N \times ((40503 \times KEY) \bmod 2^{16}) / 2^{16} \rfloor$, where KEY is the number that is fixed in this instance. There

are many other hash functions that can be used and discussions of such can be found in many textbooks dealing with computer algorithms.

[1057] In an embodiment, the message from the base station (e.g., the ECAM) is provided with a specific field, USE_OLD_SERV_CONFIG, used to indicate whether or not the parameters established in the last connection are to be used for the reconnection. This field can be used to obviate the need to send the *Service Connect Message* upon reconnection, which may reduce delay in re-establishing the connection.

[1058] Once the remote terminal has initialized the dedicated channel, it continues, for example, as described in the cdma2000 standard.

[1059] As noted above, better utilization of the reverse link resources may be achieved if the resources can be quickly allocated as needed and if available. In a wireless (and especially mobile) environment, the link conditions continually fluctuate, and long delay in allocating resources may result in inaccurate allocation and/or usage. Thus, in accordance with an aspect of the invention, mechanisms are provided to quickly assign and de-assign supplemental channels.

[1060] FIG. 4 is a diagram illustrating a communication between the remote terminal and base station to assign and de-assign a reverse link supplemental channel (R-SCH), in accordance with an embodiment of the invention. The R-SCH may be quickly assigned and de-assigned as needed. When the remote terminal has packet data to send that requires usage of the R-SCH, it requests the R-SCH by sending to the base station a *Supplemental Channel Request Mini Message* (SCRMM) (step 412). The SCRMM is a 5 msec message that may be sent on the R-DCCH or R-FCH. The base station receives the message and forwards it to the BSC (step 414). The request may or may not be granted. If the request is granted, the base station receives the grant (step 416) and transmits the R-SCH grant using a *Reverse Supplemental Channel Assignment Mini Message* (RSCAMM) (step 418). The RSCAMM is also a 5 msec message that may be sent on the F-FCH or F-DCCH (if allocated to the remote terminal) or on the F-DPCCH (otherwise). Once assigned, the remote terminal may thereafter transmit on the R-SCH (step 420).

[1061] Table 3 lists the fields for a specific format of the RSCAMM. In this embodiment, the RSCAMM includes 8 bits of layer 2 fields (i.e., the MSG_TYPE, ACK_SEQ, MSG_SEQ, and ACK_REQUIREMENT fields), 14 bits of layer 3 fields, and two reserved bits that are also used for padding as described in C.S0004 and C.S0005. The layer 3 (i.e., signaling layer) may be as defined in the cdma2000 standard.

Table 3

Field	Length (Bits)
MSG_TYPE	3
ACK_SEQUENCE	2
MSG_SEQUENCE	2
ACK_REQUIREMENT	1
REV_SCH_ID	1
REV_SCH_DURATION	4
REV_SCH_START_TIME	5
REV_SCH_NUM_BITS_IDX	4
RESERVED	2

[1062] When the remote terminal no longer has data to send on the R-SCH, it sends a *Resource Release Request Mini Message* (RRRMM) to the base station. If there is no additional signaling required between the remote terminal and base station, the base station responds with an *Extended Release Mini Message* (ERMM). The RRRMM and ERMM are also 5 msec messages that may be sent on the same channels used for sending the request and grant, respectively.

[1063] There are many scheduling algorithms that may be used to schedule the reverse link transmissions of remote terminals. These algorithms may tradeoff between rates, capacity, delay, error rates, and fairness (which gives all users some minimal level of services), to indicate some of the main criteria. In addition, the reverse link is subject to the power limitations of the remote terminal. In a single cell environment, the greatest capacity will exist when the smallest number of remote terminals is allowed to transmit with the highest rate that the remote terminal can support -- both in terms of capability and the ability

to provide sufficient power. However, in a multiple cell environment, it may be preferable for remote terminals near the boundary with another cell to transmit at a lower rate. This is because their transmissions cause interference into multiple cells -- not just a single cell. Another aspect that tends to maximize the reverse link capacity is to operate a high rise-over-thermal at the base station, which indicates high loading on the reverse link. It is for this reason that aspects of the invention use scheduling. The scheduling attempts to have a few number of remote terminals simultaneously transmit -- those that do transmit are allowed to transmit at the highest rates that they can support.

[1064] However, a high rise-over-thermal tends to result in less stability as the system is more sensitive to small changes in loading. It is for this reason that fast scheduling and control is important. Fast scheduling is important because the channel conditions change quickly. For instance, fading and shadowing processes may result in a signal that was weakly received at a base station suddenly becoming strong at the base station. For voice or certain data activity, the remote terminal autonomously changes the transmission rate While scheduling may be able to take some of this into account, scheduling may not be able to react sufficiently fast enough. For this reason, aspects of the invention provide fast power control techniques, which are described in further detail below.

[1065] An aspect of the invention provides a reliable acknowledgment/negative acknowledgment scheme to facilitate efficient and reliable data transmission. As described above, acknowledgments (Ack) and negative acknowledgments (Nak) are sent by the base station for data transmission on the R-SCH. The Ack/Nak can be sent using the F-CPANCH.

[1066] Table 4 shows a specific format for an Ack/Nak message. In this specific embodiment, the Ack/Nak message includes 4 bits that are assigned to four reverse link channels - the R-FCH, R-DCCH, R-SCH1, and R-SCH2. In an embodiment, an acknowledgment is represented by a bit value of zero ("0") and a negative acknowledgment is represented by a bit value of one ("1"). Other Ack/Nak message formats may also be used and are within the scope of the invention.

Table 4

Description	All Channels Used Number_Type (binary)	R-FCH, R-DCCH, and R-SCH1 Used Number_Type (binary)	R-FCH and R-DCCH Used Number_Type (binary)
ACK_R-FCH	xxx0	xxx0	xx00
NAK_R-FCH	xxx1	xxx1	xx11
ACK_R-DCCH	xx0x	xx0x	-
NAK_R-DCCH	xx1x	xx1x	-
ACK_R-SCH1	x0xx	00xx	00xx
NAK_R-SCH1	x1xx	11xx	11xx
ACK_R-SCH2	0xxx	-	-
NAK_R-SCH2	1xxx	-	-

[1067] In an embodiment, the Ack/Nak message is sent block coded but a CRC is not used to check for errors. This keeps the Ack/Nak message short and further allows the message to be sent with a small amount of energy. However, no coding may also be used for the Ack/Nak message, or a CRC may be attached to the message, and these variations are within the scope of the invention. In an embodiment, the base station sends an Ack/Nak message corresponding to each frame in which the remote terminal has been given permission to transmit on the R-SCH, and does not send Ack/Nak messages during frames that the remote terminal is not given permission to transmit.

[1068] During a packet data transmission, the remote terminal monitors the F-CPANCH for Ack/Nak messages that indicate the results of the transmission. The Ack/Nak messages may be transmitted from any number of base stations in the remote terminal's Active Set (e.g., from one or all base stations in the Active Set). The remote terminal can perform different actions depending on the received Ack/Nak messages. Some of these actions are described below.

[1069] If an Ack is received by the remote terminal, the data frame corresponding to the Ack may be removed from the remote terminal's physical layer transmit buffer (e.g., data source 210 in FIG. 2) since the data frame was correctly received by the base station.

[1070] If a Nak is received by the remote terminal, the data frame corresponding to the Nak may be retransmitted by the remote terminal if it is still in the physical layer transmit buffer. In an embodiment, there is a one-to-one correspondence between a forward link Ack/Nak message and a transmitted reverse link data frame. The remote terminal is thus able to identify the sequence number of the data frame not received correctly by the base station (i.e., the erased frame) based on the frame in which the Nak was received. If this data frame has not been discarded by the remote terminal, it may be retransmitted at the next available time interval, which is typically the next frame.

[1071] If neither an Ack nor a Nak was received, there are several next possible actions for the remote terminal. In one possible action, the data frame is maintained in the physical layer transmit buffer and retransmitted. If the retransmitted data frame is then correctly received at the base station, then the base station transmits an Ack. Upon correct receipt of this Ack, the remote terminal discards the data frame. This would be the best approach if the base station did not receive the reverse link transmission.

[1072] Another possible action is for the remote terminal to discard the data frame if neither an Ack nor a Nak was received. This would be the best alternative if the base station had received the frame but the Ack transmission was not received by the remote terminal. However, the remote terminal does not know the scenario that occurred and a policy needs to be chosen. One policy would be to ascertain the likelihood of the two events happening and performing the action that maximizes the system throughput.

[1073] In an embodiment, each Ack/Nak message is retransmitted a particular time later (e.g., at the next frame) to improve reliability of the Ack/Nak. Thus, if neither an Ack nor a Nak was received, the remote terminal combines the retransmitted Ack/Nak with the original Ack/Nak. Then, the remote terminal can proceed as described above. And if the combined Ack/Nak still does not result in a valid Ack or Nak, the remote terminal may discard the data frame and continue to transmit the next data frame in the sequence. The second transmission of the Ack/Nak may be at the same or lower power level relative to that of the first transmission.

[1074] If the base station did not actually receive the data frame after retransmissions, then a higher signaling layer at the base station may generate a message (e.g., an RLP NAK), which may result in the retransmission of the entire sequence of data frames that includes the erased frame.

[1075] FIG. 5A is a diagram illustrating a data transmission on the reverse link (e.g., the R-SCH) and an Ack/Nak transmission on the forward link. The remote terminal initially transmits a data frame, in frame k, on the reverse link (step 512). The base station receives and processes the data frame, and provides the demodulated frame to the BSC (step 514). If the remote terminal is in soft handoff, the BSC may also receive demodulated frames for the remote terminal from other base stations.

[1076] Based on the received demodulated frames, the BSC generates an Ack or a Nak for the data frame. The BSC then sends the Ack/Nak to the base station(s) (step 516), which then transmit the Ack/Nak to the remote terminal during frame k+1 (step 518). The Ack/Nak may be transmitted from one base station (e.g., the best base station) or from a number base stations in the remote terminal's Active Set. The remote terminal receives the Ack/Nak during frame k+1. If a Nak is received, the remote terminal retransmits the erased frame at the next available transmission time, which in this example is frame k+2 (step 520). Otherwise, the remote terminal transmits the next data frame in the sequence.

[1077] FIG. 5B is a diagram illustrating a data transmission on the reverse link and a second transmission of the Ack/Nak message. The remote terminal initially transmits a data frame, in frame k, on the reverse link (step 532). The base station receives and processes the data frame, and provides the demodulated frame to the BSC (step 534). Again, for soft handoff, the BSC may receive other demodulated frames for the remote terminal from other base stations.

[1078] Based on the received demodulated frames, the BSC generates an Ack or a Nak for the frame. The BSC then sends the Ack/Nak to the base station(s) (step 536), which then transmit the Ack/Nak to the remote terminal during frame k+1 (step 538). In this example, the remote terminal does not receive the Ack/Nak transmitted during frame k+1. However, the Ack/Nak for

the data frame transmitted in frame k is transmitted a second time during frame k+2, and is received by the remote terminal (step 540). If a Nak is received, the remote terminal retransmits the erased frame at the next available transmission time, which in this example is frame k+3 (step 542). Otherwise, the remote terminal transmits the next data frame in the sequence. As shown in FIG. 5B, the second transmission of the Ack/Nak improves the reliability of the feedback, and can result in improved performance for the reverse link.

[1079] In an alternative embodiment, the data frames are not sent back to the BSC from the base station, and the Ack/Nak is generated from the base station.

[1080] FIG. 6A is a diagram illustrating an acknowledgment sequencing with short acknowledgment delay. The remote terminal initially transmits a data frame with a sequence number of zero, in frame k, on the reverse link (step 612). For this example, the data frame is received in error at the base station, which then sends a Nak during frame k+1 (step 614). The remote terminal also monitors the F-CPANCH for an Ack/Nak message for each data frame transmitted on the reverse link. The remote terminal continues to transmit a data frame with a sequence number of one in frame k+1 (step 616).

[1081] Upon receiving the Nak in frame k+1, the remote terminal retransmits the erased frame with the sequence number of zero, in frame k+2 (step 618). The data frame transmitted in frame k+1 was received correctly, as indicated by an Ack received during frame k+2, and the remote terminal transmits a data frame with a sequence number of two in frame k+3 (step 620). Similarly, the data frame transmitted in frame k+2 was received correctly, as indicated by an Ack received during frame k+3, and the remote terminal transmits a data frame with a sequence number of three in frame k+4 (step 622). In frame k+5, the remote terminal transmits a data frame with a sequence number of zero for a new packet (step 624).

[1082] FIG. 6B is a diagram illustrating an acknowledgment sequencing with long acknowledgment delay such as when the remote terminal demodulates the Ack/Nak transmission based upon the retransmission of the Ack/Nak as described above. The remote terminal initially transmits a data frame with a sequence number of zero, in frame k, on the reverse link (step 632). The data

frame is received in error at the base station, which then sends a Nak (step 634). For this example, because of the longer processing delay, the Nak for frame *k* is transmitted during frame *k*+2. The remote terminal continues to transmit a data frame with a sequence number of one in frame *k*+1 (step 636) and a data frame with a sequence number of two in frame *k*+2 (step 638).

[1083] For this example, the remote terminal receives the Nak in frame *k*+2, but is not able to retransmit the erased frame at the next transmission interval. Instead, the remote terminal transmits a data frame with a sequence number of three in frame *k*+3 (step 640). At frame *k*+4, the remote terminal retransmits the erased frame with the sequence number of zero (step 642) since this frame is still in the physical layer buffer. Alternatively, the retransmission may be in frame *k*+3. And since the data frame transmitted in frame *k*+1 was received correctly, as indicated by an Ack received during frame *k*+3, and the remote terminal transmits a data frame with a sequence number of zero for a new packet (step 644).

[1084] As shown in FIG. 6B, the erased frame may be retransmitted at any time as long as it is still available in the buffer and there is no ambiguity as to which higher layer packet the data frame belongs to. The longer delay for the retransmission may be due to any number of reasons such as (1) longer delay to process and transmit the Nak, (2) non-detection of the first transmission of the Nak, (3) longer delay to retransmit the erased frame, and others.

[1085] An efficient and reliable Ack/Nak scheme can improve the utilization of the reverse link. A reliable Ack/Nak scheme may also allow data frames to be transmitted at lower transmit power. For example, without retransmission, a data frame needs to be transmitted at a higher power level (P_1) required to achieve one percent frame error rate (1% FER). If retransmission is used and is reliable, a data frame may be transmitted at a lower power level (P_2) required to achieve 10% FER. The 10% erased frames may be retransmitted to achieve an overall 1% FER for the transmission. Typically, $1.1 \cdot P_2 < P_1$, and less transmit power is used for a transmission using the retransmission scheme. Moreover, retransmission provides time diversity, which may improve performance. The retransmitted frame may also be combined with the first transmission of the frame at the base station, and the combined power from the two transmissions

may also improve performance. The recombining may allow an erased frame to be retransmitted at a lower power level.

[1086] An aspect of the invention provides various power control schemes for the reverse link. In an embodiment, reverse link power control is supported for the R-FCH, R-SCH, and R-DCCH. This can be achieved via a (e.g., 800 bps) power control channel, which may be partitioned into a number of power control subchannels. For example, a 100 bps power control subchannel may be defined and used for the R-SCH. If the remote terminal has not been allocated a F-FCH or F-DCCH, then the F-CPCCH may be used to send power control bits to the remote terminal.

[1087] In one implementation, the (e.g., 800 bps) power control channel is used to adjust the transmit power of the reverse link pilot. The transmit power of the other channels (e.g., the R-FCH) is set relative to that of the pilot (i.e., by a particular delta). Thus, the transmit power for all reverse link channels may be adjusted along with the pilot. The delta for each non-pilot channel may be adjusted by signaling. This implementation does not provide flexibility to quickly adjust the transmit power of different channels.

[1088] In one embodiment, the forward common power control channel (F-CPCCH) may be used to form one or more power control subchannels that may then be used for various purposes. Each power control subchannel may be defined using a number of available bits in the F-CPCCH (e.g., the m^{th} bit in each frame). For example, some of the available bits in the F-CPCCH may be allocated for a 100 bps power control subchannel for the R-SCH. This R-SCH power control subchannel may be assigned to the remote terminal during channel assignment. The R-SCH power control subchannel may then be used to (more quickly) adjust the transmit power of the designated R-SCH, e.g., relative to that of the pilot channel. For a remote terminal in soft handoff, the R-SCH power control may be based on the OR-of-the-downs rule, which decreases the transmit power if any base station in the remote terminal's Active Set directs a decrease. Since the power control is maintained at the base station, this permits the base station to adjust the transmitted power with minimal amount of delay and thus adjust the loading on the channel.

[1089] The R-SCH power control subchannel may be used in various manners to control the transmission on the R-SCH. In an embodiment, the R-SCH power control subchannel may be used to direct the remote terminal to adjust the transmit power on the R-SCH by a particular amount (e.g., 1 dB, 2 dB, or some other value). In another embodiment, the subchannel may be used to direct the remote terminal to reduce or increase transmit power by a large step (e.g., 3 dB, or possibly more). In both embodiments, the adjustment in transmit power may be relative to the pilot transmit power. In another embodiment, the subchannel may be directed to adjust the data rate allocated to the remote terminal (e.g., to the next higher or lower rate). In yet another embodiment, the subchannel may be used to direct the remote terminal to temporarily stop transmission. And in yet another embodiment, the remote terminal may apply different processing (e.g., different interleaving interval, different coding, and so on) based on the power control command. The R-SCH power control subchannel may also be partitioned into a number of "sub-subchannels", each of which may be used in any of the manners described above. The sub-subchannels may have the same or different bit rates. The remote terminal may apply the power control immediately upon receiving the command, or may apply the command at the next frame boundary.

[1090] The ability to reduce the R-SCH transmit power by a large amount (or down to zero) without terminating the communication session is especially advantageous to achieve better utilization of the reverse link. Temporary reduction or suspension of a packet data transmission can typically be tolerated by the remote terminal. These power control schemes can be advantageously used to reduce interference from a high rate remote terminal.

[1091] Power control of the R-SCH may be achieved in various manners. In one embodiment, a base station monitors the received power from the remote terminals with a power meter. The base station may even be able to determine the amount of power received from each channel (e.g., the R-FCH, R-DCCH, R-SCH, and so on). The base station is also able to determine the interference, some of which may be contributed by remote terminals not being served by this base station. Based on the collected information, the base station may adjust the transmit power of some or all remote terminals based on various factors.

For example, the power control may be based on the remote terminals' category of service, recent performance, recent throughput, and so on. The power control is performed in a manner to achieve the desired system goals.

[1092] Power control may be implemented in various manners. Example implementations are described in U.S Patent No. 5,485,486, entitled "METHOD AND APPARATUS FOR CONTROLLING TRANSMISSION POWER IN A CDMA CELLULAR MOBILE TELEPHONE SYSTEM," issued January 16, 1996, U.S Patent No. 5,822,318, entitled "METHOD AND APPARATUS FOR CONTROLLING POWER IN A VARIABLE RATE COMMUNICATION SYSTEM," issued October 13, 1998, and U.S Patent No. 6,137,840, entitled "METHOD AND APPARATUS FOR PERFORMING FAST POWER CONTROL IN A MOBILE COMMUNICATION SYSTEM," issued October 24, 2000, all assigned to the assignee of the present application and incorporated herein by reference.

[1093] In a typical method of power control that is used to control the level of the R-PICH channel, the base station measures the level of the R-PICH, compares it to a threshold, and then determines whether to increase or decrease the power of the remote terminal. The base station transmits a bit to the remote terminal instructing it to increase or decrease its output power. If the bit is received in error, the remote terminal will transmit at the incorrect power. During the next measurement of the R-PICH level received by the base station, the base station will determine that the received level is not at the desired level and send a bit to the remote terminal to change its transmit power. Thus, bit errors do not accumulate and the loop controlling the remote terminal's transmit power will stabilize to the correct value.

[1094] Errors in the bits sent to the remote terminal to control the traffic-to-pilot ratio for congestion power control can cause the traffic-to-pilot ratio to be other than that desired. However, the base station typically monitors the level of the R-PICH for reverse power control or for channel estimation. The base station can also monitor the level of the received R-SCH. By taking the ratio of the R-SCH level to the R-PICH level, the base station can estimate the traffic-to-pilot ratio in use by the remote terminal. If the traffic-to-pilot ratio is not that which is desired, then the base station can set the bit that controls the traffic-to-

pilot ratio to correct for the discrepancy. Thus, there is a self-correction for bit errors.

[1095] Once a remote terminal has received a grant for the R-SCH, the remote terminal typically transmits at the granted rate (or below in case it doesn't have enough data to send or does not have sufficient power) for the duration of the grant. The channel load from other remote terminals can vary quite quickly as a result of fading and the like. As such, it may be difficult for the base station to estimate the loading precisely in advance.

[1096] In an embodiment, a "congestion" power control subchannel may be provided to control a group of remote terminals in the same manner. In this case, instead of a single remote terminal monitoring the power control subchannel to control the R-SCH, a group of remote terminals monitor the control subchannel. This power control subchannel can be at 100 bps or at any other transmission rate. In one embodiment, the congestion control subchannel is implemented with the power control subchannel used for the R-SCH. In another embodiment, the congestion control subchannel is implemented as a "sub-subchannel" of the R-SCH power control subchannel. In yet another embodiment, the congestion control subchannel is implemented as a subchannel different from the R-SCH power control subchannel. Other implementations of the congestion control subchannel may also be contemplated and are within the scope of the invention.

[1097] The remote terminals in the group may have the same category service (e.g., remote terminals having low priority available bit rate services) and may be assigned to a single power control bit per base station. This group control based on a single power control stream performs similar to that directed to a single remote terminal to provide for congestion control on the reverse link. In case of capacity overload, the base station may direct this group of remote terminals to reduce their transmit power or their data rates, or to temporarily stop transmitting, based on a single control command. The reduction in the R-SCH transmit power in response to the congestion control command may be a large downward step relative to the transmit power of the pilot channel.

[1098] The advantage of a power control stream going to a group of remote terminals instead of a single remote terminal is that less overhead power is

required on the forward link to support the power control stream. It should be noted that the transmit power of a bit in the power control stream can be equal to the power of the normal power control stream used to the control the pilot channel for the remote terminal that requires the most power. That is, the base station can determine the remote terminal in the group that requires the greatest power in its normal power control stream and then use this power to transmit the power control bit used for congestion control.

[1099] FIG. 7 is a flow diagram that illustrates a variable rate data transmission on the R-SCH with fast congestion control, in accordance with an embodiment of the invention. During the transmission on the R-SCH, the remote terminal transmits in accordance with the data rate granted in the *Reverse Supplemental Channel Assignment Mini Message* (RSAMM). If variable rate operation is permitted on the R-SCH, the remote terminal may transmit at any one of a number of permitted data rates.

[1100] If the remote terminal's R-SCH has been assigned to a congestion control subchannel, then, in an embodiment, the remote terminal adjusts the traffic-to-pilot ratio based upon the bits received in the congestion control subchannel. If variable rate operation is permitted on the R-SCH, the remote terminal checks the current traffic-to-pilot ratio. If it is below the level for a lower data rate, then the remote terminal reduces its transmission rate to the lower rate. If it is equal to or above the level for a higher data rate, then the remote terminal increases its transmission rate to the higher rate if it has sufficient data to send.

[1101] Prior to the start of each frame, the remote terminal determines the rate to use for transmitting the next data frame. Initially, the remote terminal determines whether the R-SCH traffic-to-pilot ratio is below that for the next lower rate plus a margin Δ_{low}, at step 712. If the answer is yes, a determination is made whether the service configuration allows for a reduction in the data rate, at step 714. And if the answer is also yes, the data rate is decreased, and the same traffic-to-pilot ratio is used, at step 716. And if the service configuration does not allow for a rate reduction, a particular embodiment would permit the remote terminal to temporarily stop transmitting.

[1102] Back at step 712, if the R-SCH traffic-to-pilot ratio is not above that for the next lower data rate plus the margin Δ_{low}, a determination is next made as to whether the R-SCH traffic-to-pilot ratio is greater than that for the next higher data rate minus a margin Δ_{high}, at step 718. If the answer is yes, a determination is made whether the service configuration allows for an increase in the data rate, at step 720. And if the answer is also yes, the transmission rate is increased, and the same traffic-to-pilot ratio is used, at step 722. And if the service configuration does not allow for a rate increase, the remote terminal transmits at the current rate.

[1103] FIG. 8 is a diagram illustrating improvement that may be possible with fast control of the R-SCH. On the left frame, without any fast control of the R-SCH, the rise-over-thermal at the base station varies more widely, exceeding the desired rise-over-thermal level by a larger amount in some instances (which may result in performance degradation for the data transmissions from the remote terminals), and falling under desired rise-over-thermal level by a larger amount in some other instances (resulting in under-utilization of the reverse link resources). In contrast, on the right frame, with fast control of the R-SCH, the rise-over-thermal at the base station is maintained more closely to the desired rise-over-thermal level, which results in improved reverse link utilization and performance.

[1104] In an embodiment, a base station may schedule more than one remote terminal (via SCAM or ESCAM) to transmit, in response to receiving multiple requests (via SCRM or SCRMM) from different remote terminals. The granted remote terminals may thereafter transmit on the R-SCH. If overloading is detected at the base station, a "fast reduce" bit stream may be used to turn off (i.e., disable) a set of remote terminals (e.g., all except one remote terminal). Alternatively, the fast reduce bit stream may be used to reduce the data rates of the remote terminals (e.g., by half). Temporarily disabling or reducing the data rates on the R-SCH for a number of remote terminals may be used for congestion control, as described in further detail below. The fast reduce capability may also be advantageously used to shorten the scheduling delay.

[1105] When the remote terminals are not in soft handoff with other base stations, the decision on which remote terminal is the most advantaged (efficient) to use the reverse link capacity may be made at the BTS. The most efficient remote terminal may then be allowed to transmit while the others are temporarily disabled. If the remote terminal signals the end of its available data, or possibly when some other remote terminal becomes more efficient, the active remote terminal can quickly be changed. These schemes may increase the throughput of the reverse link.

[1106] In contrast, for a usual set up in a cdma2000 system, a R-SCH transmission can only start or stop via layer 3 messaging, which may take several frames from composing to decoding at the remote terminal to get across. This longer delay causes a scheduler (e.g., at the base station or BSC) to work with (1) less reliable, longer-term predictions about the efficiency of the remote terminal's channel condition (e.g., the reverse link target pilot Ec/(No+Io) or set point), or (2) gaps in the reverse link utilization when a remote terminal notifies the base station of the end of its data (a common occurrence since a remote terminal often claims it has a large amount of data to send to the base station when requesting the R-SCH).

[1107] Referring back to FIG. 2, the elements of remote terminal 106 and base station 104 may be designed to implement various aspects of the invention, as described above. The elements of the remote terminal or base station may be implemented with a digital signal processor (DSP), an application specific integrated circuit (ASIC), a processor, a microprocessor, a controller, a microcontroller, a field programmable gate array (FPGA), a programmable logic device, other electronic units, or any combination thereof. Some of the functions and processing described herein may also be implemented with software executed on a processor, such as controller 230 or 270.

[1108] Headings are used herein to serve as general indications of the materials being disclosed, and are not intended to be construed as to scope.

[1109] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those

skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1110] WHAT IS CLAIMED IS:

CLAIMS

1. A channel structure capable of supporting data transmission on a reverse link of a wireless communication system, comprising:

a reverse fundamental channel configurable to transmit data and signaling on the reverse link;

a reverse supplemental channel assignable and configurable to transmitted packet data on the reverse link;

a reverse control channel configurable to transmit signaling on the reverse link; and

a forward power control channel configurable to transmit first and second power control streams for the reverse link for a particular remote terminal, wherein

the first power control stream is used to control the transmit power of the reverse supplemental channel in combination with at least one other reverse link channel, and

the second power control stream is used to control a transmit characteristic of the reverse supplemental channel.

2. The channel structure of claim 1, wherein the second power control stream is used to control the transmit power of the reverse supplemental channel relative to that of a designated reverse link channel.

3. The channel structure of claim 1, wherein the second power control stream is used to control the data rate of the reverse supplemental channel.

4. The channel structure of claim 1, further comprising:

a forward acknowledgment channel configurable to transmit, on the forward link, signaling indicative of received status of the packet data transmission on the reverse link.

5. The channel structure of claim 4, wherein the forward acknowledgment channel is configurable to transmit an acknowledgment or a negative acknowledgment for each transmitted data frame on the reverse supplemental channel.

6. The channel structure of claim 5, wherein the acknowledgment or negative acknowledgment for each transmitted data frame is transmitted a plurality of times on the forward acknowledgment channel.

7. The channel structure of claim 1, wherein the reverse control channel is configurable to transmit signaling used to assign and de-assign the reverse supplemental channel.

8. The channel structure of claim 1, further comprising:

a reverse rate indicator channel configurable to transmit on the reverse link information related to a packet data transmission on the reverse link.

9. A channel structure capable of supporting data transmission on a reverse link of a wireless communication system, comprising:

a reverse fundamental channel configurable to transmit data and signaling on the reverse link;

a reverse supplemental channel assignable and configurable to transmitted packet data on the reverse link;

a reverse control channel configurable to transmit signaling on the reverse link; and

a forward power control channel configurable to transmit first and second power control streams for the reverse link for a particular remote terminal, wherein

the first power control stream is used to control the transmit power of the reverse supplemental channel in combination with at least one other reverse link channel, and

the second power control stream is configured to control a transmit characteristic of a group of remote terminals.

10. The channel structure of claim 9, wherein the second power control stream is used to similarly control the transmit power or data rate of the group of remote terminals.

11. The channel structure of claim 9, wherein the second power control stream is used to enable and disable transmissions on reverse supplemental channels assigned to the group of remote terminals.

12. A method for transmitting data on a reverse link of a wireless communication system, comprising:

transmitting a frame of data on the reverse link via a data channel;

temporarily retaining the data frame in a buffer;

monitoring for a message on a forward link indicating a received status of the transmitted data frame; and

processing the data frame based on the received message.

13. The method of claim 12, wherein the processing includes;

retransmitting the data frame if the message indicates that the transmitted data frame was incorrectly received.

14. The method of claim 12, wherein the processing includes;

discarding the data frame from the buffer if the message indicates that the transmitted data frame was correctly received.

15. The method of claim 12, wherein the processing includes;

retaining the data frame in the buffer if the message is not properly detected.

16. The method of claim 12, further comprising:

monitoring for a second transmission of the message;

wherein the processing of the data frame is based on one or more received messages for the data frame.

17. The method of claim 16, further comprising:
combining the received messages for the data frame to provide a more reliable message.

18. The method of claim 12, further comprising:
identifying the transmitted data frame with a sequence number.

19. The method of claim 18, further comprising:
transmitting the sequence number of the transmitted data frame via a signaling channel.

20. The method of claim 12, further comprising:
identifying the transmitted data frame as either a first transmission or a retransmission.

21. A method for transmitting data on a reverse link of a wireless communication system, comprising:
transmitting a frame of data on the reverse link via a data channel;
temporarily retaining the data frame in a buffer;
monitoring for a message on a forward link indicating a received status of the transmitted data frame;
retransmitting the data frame if the message indicates that the transmitted data frame was incorrectly received;
discarding the data frame from the buffer if the message indicates that the transmitted data frame was correctly received; and
retaining the data frame in the buffer if the message is not properly detected.

22. A method for controlling transmit power of a supplemental channel in a reverse link of a wireless communication system, comprising:

receiving a first power control stream for controlling the transmit power of the supplemental channel in combination with at least one other reverse link channel;

receiving a second power control stream for controlling a transmit characteristic of the supplemental channel; and

adjusting the transmit power and characteristic of the supplemental channel based on the first and second power control streams.

23. The method of claim 22, wherein the second power control stream controls the transmit power of the supplemental channel relative to that of a designated reverse link channel.

24. The method of claim 22, wherein the second power control stream controls a data rate of the supplemental channel.

25. The method of claim 22, wherein the second power control stream enables and disables transmission on the supplemental channel.

26. The method of claim 22, wherein the transmit power of the supplemental channel is adjusted by a larger step in response to the second power control stream than for the first power control stream.

27. The method of claim 22, wherein the second power control stream is assigned to a plurality of remote terminals.

28. The method of claim 28, wherein supplemental channels for the plurality of remote terminals are controlled in similar manner by the second power control stream.

29. A remote terminal in a wireless communication system, comprising:

a transmit data processor configurable to process and transmit

data and signaling on a reverse fundamental channel,

packet data on an assigned reverse supplemental channel,

signaling on a reverse control channel, and

information related to a packet data transmission on a reverse indicator channel;

a receive data processor configurable to receive a plurality of power control streams on a forward power control channel; and

a controller operatively coupled to the transmit and receive data processors and configured to control one or more transmit characteristics of the reverse supplemental channel based on the plurality of power control streams.

30. The remote terminal of claim 29, wherein the receive data processor is further configurable to receive, on a forward acknowledgment channel, signaling indicative of received status of a packet data transmission on the reverse supplemental channel.

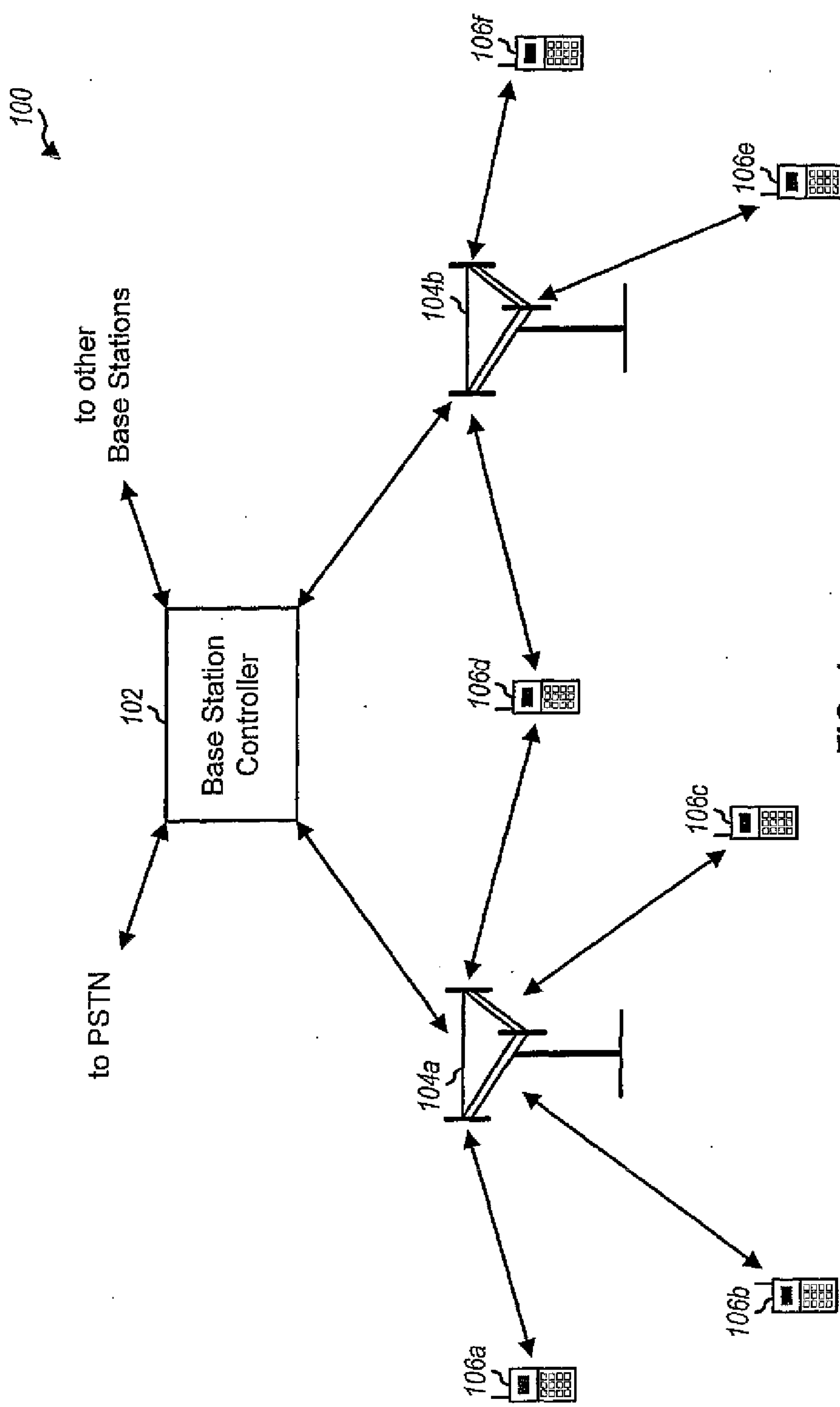

100
to other
Base Stations
102
Base Station
Controller
to PSTN
104a
104b
106a
106b
106c
106d
106e
106f

FIG. 1

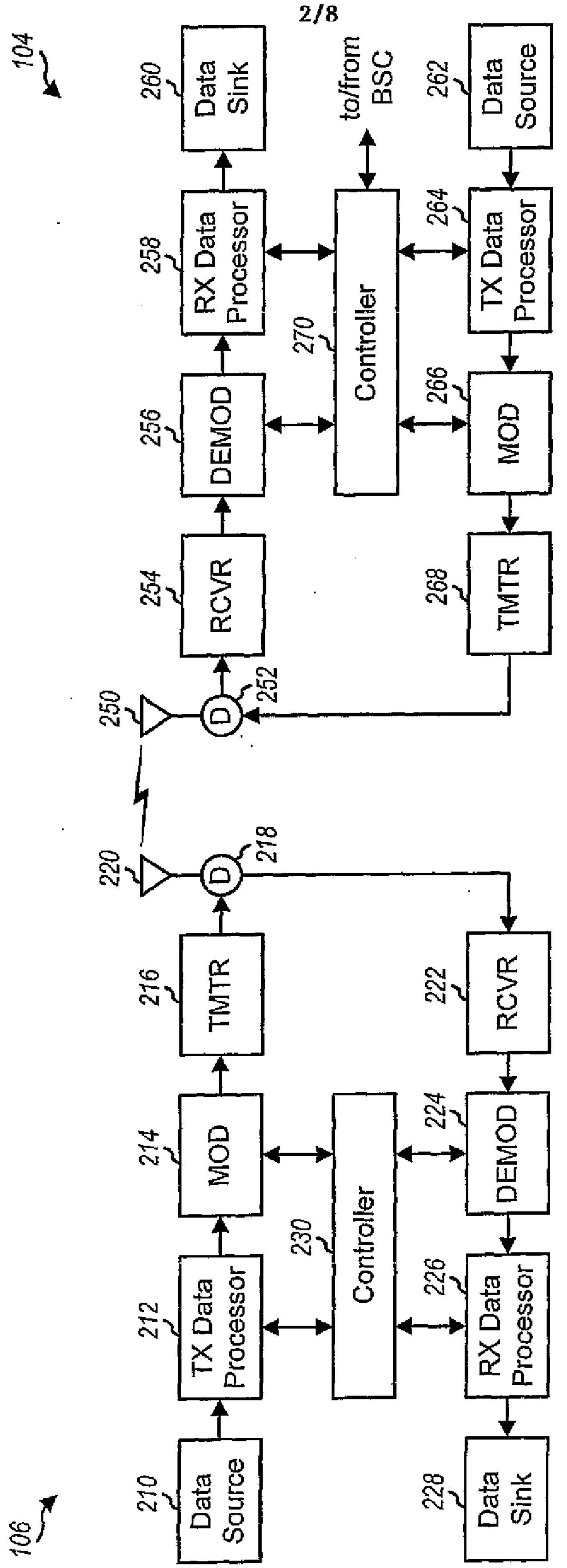

104
260 Data Sink
258 RX Data Processor
256 DEMOD
254 RCVR
252 D
250
to/from BSC
270 Controller
262 Data Source
264 TX Data Processor
266 MOD
268 TMTR
220
218 D
216 TMTR
214 MOD
212 TX Data Processor
210 Data Source
230 Controller
222 RCVR
224 DEMOD
226 RX Data Processor
228 Data Sink
106

FIG. 2

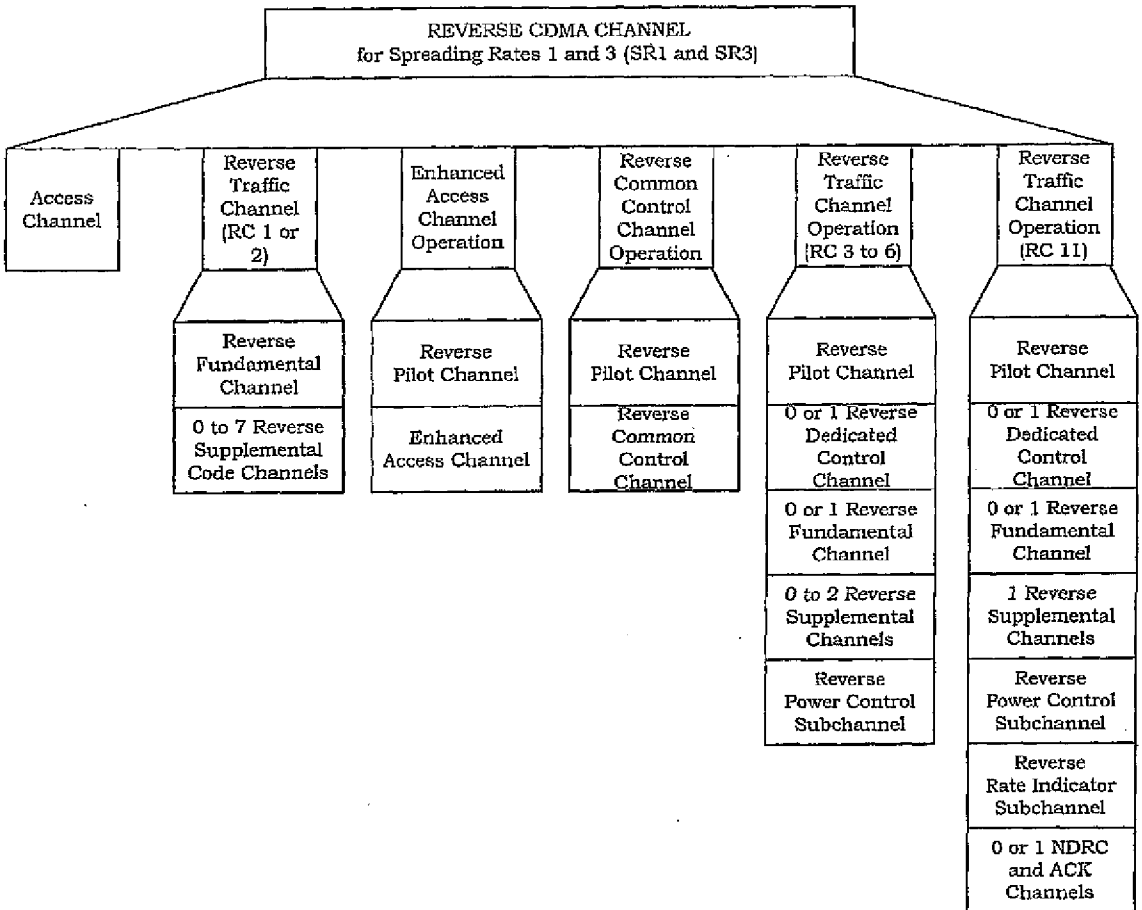

REVERSE CDMA CHANNEL for Spreading Rates 1 and 3 (SR1 and SR3)
Access Channel
Reverse Traffic Channel (RC 1 or 2)
Enhanced Access Channel Operation
Reverse Common Control Channel Operation
Reverse Traffic Channel Operation (RC 3 to 6)
Reverse Traffic Channel Operation (RC 11)
Reverse Fundamental Channel
0 to 7 Reverse Supplemental Code Channels
Reverse Pilot Channel
Enhanced Access Channel
Reverse Pilot Channel
Reverse Common Control Channel
Reverse Pilot Channel
0 or 1 Reverse Dedicated Control Channel
0 or 1 Reverse Fundamental Channel
0 to 2 Reverse Supplemental Channels
Reverse Power Control Subchannel
Reverse Pilot Channel
0 or 1 Reverse Dedicated Control Channel
0 or 1 Reverse Fundamental Channel
1 Reverse Supplemental Channels
Reverse Power Control Subchannel
Reverse Rate Indicator Subchannel
0 or 1 NDRC and ACK Channels

FIG. 3A

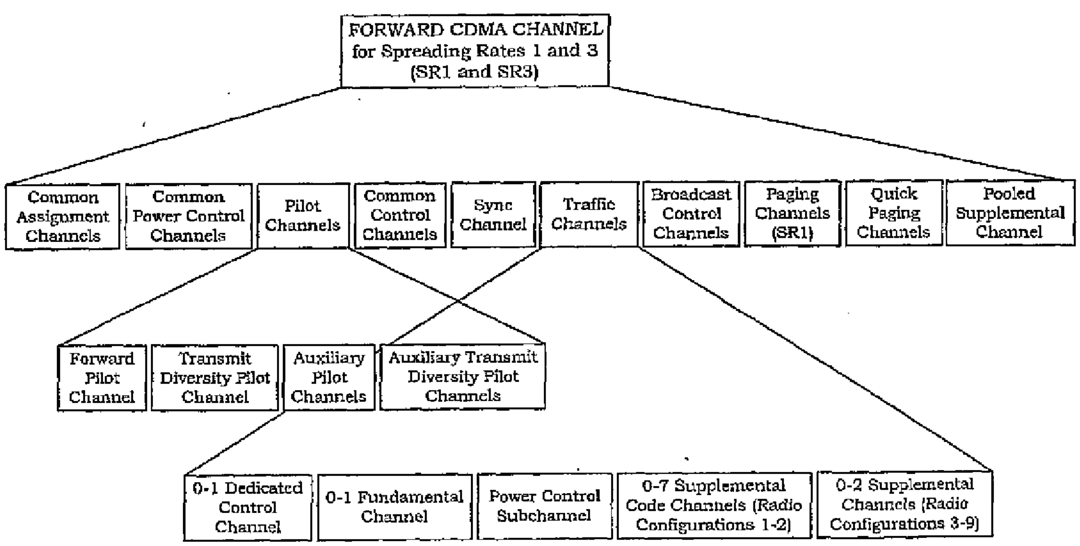

FORWARD CDMA CHANNEL for Spreading Rates 1 and 3 (SR1 and SR3)
Common Assignment Channels
Common Power Control Channels
Pilot Channels
Common Control Channels
Sync Channel
Traffic Channels
Broadcast Control Channels
Paging Channels (SR1)
Quick Paging Channels
Pooled Supplemental Channel
Forward Pilot Channel
Transmit Diversity Pilot Channel
Auxiliary Pilot Channels
Auxiliary Transmit Diversity Pilot Channels
0-1 Dedicated Control Channel
0-1 Fundamental Channel
Power Control Subchannel
0-7 Supplemental Code Channels (Radio Configurations 1-2)
0-2 Supplemental Channels (Radio Configurations 3-9)

FIG. 3B